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SUMMARY TECHNICAL REPORT
OF THE
NATIONAL DEFENSE RESEARCH COMMITTEE

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SUMMARY TECHNICAL REPORT OF DIVISION 13, NDRC

VOLUME 2A

COMMUNICATION RESEARCH

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VANNEVAR BUSH, DIRECTOR

NATIONAL DEFENSE RESEARCH COMMITTEE
JAMES B. CONANT, CHAIRMAN

DIVISION 13
HARADEN PRATT, CHIEF

WASHINGTON, D. C., 1946

[redacted]

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NOTES ON THE ORGANIZATION OF NDRC

The duties of the National Defense Research Committee were (1) to recommend to the Director of OSRD suitable projects and research programs on the instrumentalities of warfare, together with contract facilities for carrying out these projects and programs, and (2) to administer the technical and scientific work of the contracts. More specifically, NDRC functioned by initiating research projects on requests from the Army or the Navy, or on requests from an allied government transmitted through the Liaison Office of OSRD, or on its own considered initiative as a result of the experience of its members. Proposals prepared by the Division, Panel, or Committee for research contracts for performance of the work involved in such projects were first reviewed by NDRC, and if approved, recommended to the Director of OSRD. Upon approval of a proposal by the Director, a contract permitting maximum flexibility of scientific effort was arranged. The business aspects of the contract, including such matters as materials, clearances, vouchers, patents, priorities, legal matters, and administration of patent matters were handled by the Executive Secretary of OSRD.

Originally NDRC administered its work through five divisions, each headed by one of the NDRC members. These were:

Division A—Armor and Ordnance
Division B—Bombs, Fuels, Gases, & Chemical Problems
Division C—Communication and Transportation
Division D—Detection, Controls, and Instruments
Division E—Patents and Inventions

In a reorganization in the fall of 1942, twenty-three administrative divisions, panels, or committees were created, each with a chief selected on the basis of his outstanding work in the particular field. The NDRC members then became a reviewing and advisory group to the Director of OSRD. The final organization was as follows:

Division 1—Ballistic Research
Division 2—Effects of Impact and Explosion
Division 3—Rocket Ordnance
Division 4—Ordnance Accessories
Division 5—New Missiles
Division 6—Sub-Surface Warfare
Division 7—Fire Control
Division 8—Explosives
Division 9—Chemistry
Division 10—Absorbents and Aerosols
Division 11—Chemical Engineering
Division 12—Transportation
Division 13—Electrical Communication
Division 14—Radar
Division 15—Radio Coordination
Division 16—Optics and Camouflage
Division 17—Physics
Division 18—War Metallurgy
Division 19—Miscellaneous
Applied Mathematics Panel
Applied Psychology Panel
Committee on Propagation
Tropical Deterioration Administrative Committee

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NDRC FOREWORD

As events of the years preceding 1940 revealed more and more clearly the seriousness of the world situation, many scientists in this country came to realize the need of organizing scientific research for service in a national emergency. Recommendations which they made to the White House were given careful and sympathetic attention, and as a result the National Defense Research Committee [NDRC] was formed by Executive Order of the President in the summer of 1940. The members of NDRC, appointed by the President, were instructed to supplement the work of the Army and the Navy in the development of the instrumentalities of war. A year later, upon the establishment of the Office of Scientific Research and Development [OSRD], NDRC became one of its units.

The Summary Technical Report of NDRC is a conscientious effort on the part of NDRC to summarize and evaluate its work and to present it in a useful and permanent form. It comprises some seventy volumes broken into groups corresponding to the NDRC Divisions, Panels, and Committees.

The Summary Technical Report of each Division, Panel, or Committee is an integral survey of the work of that group. The first volume of each group's report contains a summary of the report, stating the problems presented and the philosophy of attacking them, and summarizing the results of the research, development, and training activities undertaken. Some volumes may be "state of the art" treatises covering subjects to which various research groups have contributed information. Others may contain descriptions of devices developed in the laboratories. A master index of all these divisional, panel, and committee reports which together constitute the Summary Technical Report of NDRC is contained in a separate volume, which also includes the index of a microfilm record of pertinent technical laboratory reports and reference material.

Some of the NDRC-sponsored researches which had been declassified by the end of 1945 were of sufficient popular interest that it was found desirable to report them in the form of monographs, such as the series on radar by Division 14 and the monograph on sampling inspection by the Applied Mathematics Panel. Since the material treated in them is not duplicated in the Summary Technical Report of NDRC, the monographs are an important part of the story of these aspects of NDRC research.

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In contrast to the information on radar, which is of widespread interest and much of which is released to the public, the research on subsurface warfare is largely classified and is of general interest to a more restricted group. As a consequence, the report of Division 6 is found almost entirely in its Summary Technical Report, which runs to over twenty volumes. The extent of the work of a Division cannot therefore be judged solely by the number of volumes devoted to it in the Summary Technical Report of NDRC; account must be taken of the monographs and available reports published elsewhere.

Of all the NDRC Divisions, few were larger or charged with more diverse responsibilities than Division 13. Under the urgent pressure of wartime requirements, the staff of the Division developed navigation and communications devices and systems which not only contributed to the successful Allied war effort, but which will continue to be of value in time of peace in the fields of transportation and communications. The work of the Division, under the direction first of C. B. Jolliffe and later of Haraden Pratt, furnishes a foundation for what promises to be even more radical developments than those of the war—for one example, direction finders which will operate at all elevations and azimuth angles, in other words, hemispherically.

The Summary Technical Report of Division 13 was prepared under the direction of the Division Chief and authorized by him for publication. The report presents the methods and results of the widely varied research and development program, and, in the case of work with speech scrambling and decoding, it presents for the first time a comprehensive review of the state of the art. The report is also a notable record of the skill and integrity of the scientists and engineers, who, with the cooperation of the Army and Navy and Division contractors, contributed brilliantly to the defense of the nation. To all of these we express our sincere appreciation.

Vannevar Bush, Director
Office of Scientific Research and Development

J. B. Conant, Chairman
National Defense Research Committee

FOREWORD

At the outset of the national emergency, which made it desirable to focus attention on military requirements in communications, developments in this field had reached a very advanced state in commercial applications. It was felt that because of this, and because of the existence of many excellent research and development organizations in industry, it would not be necessary to establish any special research activity or central laboratory, but only to assess through Division 13 of the National Defense Research Committee the special problems for which military requirements needed solution, and present them to these already staffed groups.

Naturally, many military needs differed radically from those encountered in civil life. There was also the urge to push ahead as fast as possible in the region of very high and ultra high frequencies where industry was feeling its way, being guided largely by the needs of commerce as such fields slowly unfolded. If the problems of generation, control, modulation, and reception in these frontier frequency areas could be harnessed, it was felt that results of inestimable value to the war effort would quickly follow.

Another factor that influenced the types of research activity was the need to implement the new and greatly enlarged role of aircraft by giving airplanes the utmost in communication and in protection from adverse electrical weather conditions, and by utilizing the mobility of aircraft to create radio interference to enemy operations.

Thus the Division adopted a very wide field of related activities and, at the outset, initiated projects of fundamental importance such as development of ultra-high-frequency generators and measuring apparatus, precipitation static research to free aircraft radio reception from blackouts due to charged particles in the air, reception methods to make easy the identification of strange radio signals, and propagation studies of radio waves on a world-wide coordinated basis. Many individual special studies in apparatus design and miscellaneous subjects were taken up as needs arose.

Recognizing the need for radio interference generators the Division initiated two projects on this subject which were carried to conclusion with successful field demonstrations. This general activity became so important that a new Division 15 was organized which took over this particular activity.

As the Division's activities progressed and the array of communication, navigation, and identification needs became more complex, it was found that the study of systems as a whole required special attention. The problem of an adequate communication network and its proper integration into the early air warning systems of the Army Air Forces loomed large and became a very important Division project. As time went on, the emphasis shifted more and more from equipments and instrumentalities to systems engineering problems. This led to the need of evaluating systems so that future planning could be done in the light of intelligent appraisal of the respective values of methods of both friend and foe.

Towards the end of the war, this shift towards system problems led to a definite need for a central laboratory. The war progress in special military communication developments had opened up new fields requiring the attention of specialists outside of existing laboratories. This brought about the establishment in 1944 of Central Communications Research at Harvard University, where a very considerable staff of workers was active under the direction of Professor Chaffee, up to the time of demobilization. Much of this program continued under direct contracts between Harvard University and the military services.

Haraden Pratt
Chief, Division 13

PREFACE

In summarizing the several hundred reports of contractors on the hundred-odd research projects sponsored by Division 13 of the National Defense Research Committee [NDRC], the editor has had to settle in his own mind how much or how little of each project report should be included; in other words, how far the boiling-down process should go.

The editor has an abhorrence for seeing good scientific or technical material go unpublished. Only by publication can the facts or methods developed by a few researchers become available for all researchers. On this basis, substantially all Division 13's program should be included in the volumes, of which this is one, summarizing the work of the Division. On the other hand, time moves forward inexorably so that it is quite likely that, by the day of publication, much of the data would already be out of date. Furthermore, time and human energy are always scarce. On these bases, all that might be required would be a paragraph or two summarizing the aims of the project and its accomplishments.

A middle course was steered, a course between the easiest solution of publishing practically all of each report and the more difficult job of really digesting the project purpose and results. The editor, however, deliberately chose to publish too much rather than too little. In most cases it will be unnecessary for the reader to search out the original source material unless he wishes to dig deep into the subject. In those cases where fundamental information was assembled and printed in the project report, that is, information on which future research might be based, the summaries have been permitted to take as much space as required.

This particular volume contains summaries of projects dealing with a very wide variety of subjects and illustrates the editorial technique.

Among many reports dealing with a whole series of projects on panoramic reception, the editor found a concise thesis called *The Fundamentals of Panoramic Reception*. This report is basic; all present and future schemes of panoramic reception draw upon the facts in it. The greater part of this report appears in this volume. On the other hand, only brief descriptions of actual apparatus developed by the Division using these principles are found here.

Early in the life of the Division it was realized that the microwave region of the ether spectrum would be extensively used during the war. At that time very little was known of the propagation characteristics, or, in fact, of how to build microwave apparatus. Equipment was designed, built, and field-tested at frequencies of the order of 3,000 megacycles. Out of this came an omnidirectional system useful as an additional channel of communication; a highly portable transmitter-receiver with searchlight directivity; and the first field-strength measuring equipment for 3,000 megacycles. Near the end of World War II, numerous systems were in operation or proposed for frequencies above 1,000 megacycles. An extensive analysis of all these systems was made by the Division. All this work is summarized in this volume.

Another important subject covered here is the work done toward discovering the causes of and alleviating the effects of airplane precipitation static. Part of the Division's work led to the "block and squirter" system for reducing radio interference; another part disclosed the fact that uncoated airplane surfaces pick up less static than those which have been painted. An important end result was the acquisition of exact information on structural changes to airplanes and their appurtenances, such as antennas, needed to avoid the generation of corona discharges except at controlled points.

One of the biggest jobs of the Division was to aid the Air Forces in designing, procuring, installing, and operating airplane warning systems abroad. Under the general subject of "systems" research, each of the component parts was studied, and all these parts were properly integrated into a system. In fact, the Division was forced to aid in packaging some of the components so that they arrived in working condition at the European Theater of Operations! So vast was the work done under this general subject that only the merest summary could be included in this volume.

A great amount of cooperative work on studying the vagaries of the ionosphere, with particular reference to the use of high frequencies for direction finding, was coordinated by the Division, all aimed at the production of a service prognosticating radio transmission conditions on a world-wide basis. This work was finally taken over by the Interdepartmental Radio Propagation Laboratory [IRPL] and continues under sponsorship of the National Bureau of Standards.

Other topics studied were the effects of trees and hills on radio communication in the region of 4 to 116 megacycles; means of shielding diathermy machines to reduce or prevent radio interference; the possibility of growing frequency-standard crystals to augment the natural sources of quartz; the development of a remarkably stable oscillator which needed no crystal at all; reconnaissance television; airplane facsimile equipment; the existing means of recording sounds on magnetic materials; the psychological bases on which effective jamming of enemy radio signals could occur—including the design and construction of two sets of jamming apparatus. Jamming and countermeasures became of such vital importance that a new Division (15) of NDRC was organized to carry forward at an accelerated pace the early work of Division 13.

Among the noteworthy accomplishments of the Division are the production of a means of "flash" telegraphy by which short radio messages can be transmitted and received at the rate of 3,000 to 9,000 words per minute; a new method of locating faults in field wire, and an elegant method of laying such wire by airplane; and the development of new batteries which were useful at temperatures as low as —40 F.

Finally, the reader's attention is directed to material supplementing this volume, which is published as Volume 2B of the Division 13 Summary Technical Report, *Electronic Navigation Systems.*

Keith Henney
Editor

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CONTENTS

PART I
GENERAL COMMUNICATIONS RESEARCH

CHAPTER PAGE
1 Systems Engineering for AAF Communications . . . 3
2 Propagation Studies 5
3 Frequency Modulation Versus Amplitude Modulation . 8
4 Microwave Communication Systems 11

PART II
3,000-Mc COMMUNICATION

5 Field Tests and 3,000-Mc Equipment 19
6 R-F Generator for 2,000 to 3,000 Mc 37

PART III
PRECIPITATION STATIC PROBLEMS 45
7 Precipitation Static Reduction 47
8 Precipitation Static Research 49
9 Effect of Aircraft Surface Treatment 51
10 Noise Eliminator Tests 53
11 The Block-and-Squirter System 57

PART IV
PANORAMIC RECEPTION 61
12 The Fundamentals of Panoramic Reception 63
13 Panoramic Receiver with Moving-Screen Indicator . . 88
14 Improved Panoramic Receiver 98
15 Receiver for Pulse Signals 108

PART V
INTERFERENCE GENERATION 117
16 Study in Interference Generation 119
17 Radio Interference Generators 122

PART VI
RADIO TRANSMISSION FORECASTING

18 Ionosphere Studies 131

[redacted]

PART VII
APPARATUS DESIGN

CHAPTER PAGE
19 U-H-F Field-Intensity Measuring Equipment 137
20 Aircraft Facsimile Systems 141
21 Ultra-High-Speed Flash Telegraphy 144
22 Frequency-Stabilized Master Oscillator 154
23 Pickup Tube for Reconnaissance Television. 162
24 Sound Recording on Magnetic Materials 166

PART VIII
MISCELLANEOUS STUDIES

25 Substitutes for Natural Quartz for Frequency Control . 173
26 Shielding for Diathermy 175
27 Locating Faults in Wire Lines 176
28 Storage Batteries for Cold Climates 181
29 An Electrical Cancellation and Indicating System . . 183
30 Laying Field Wire by Airplane 191
31 Floating Insulated Wire 192
32 Noise Investigations 193
33 Transient Response of Band-Pass Amplifiers 196
34 Measurements of Magnetic Properties of Ferrite Core Materials 198
35 Aircraft Antenna Power, Impedance, and Tuning-Network Survey 200
Bibliography 203
OSRD Appointees 208
Contract Numbers 209
Project Numbers 211
Index 213

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PART I

GENERAL COMMUNICATIONS RESEARCH

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Chapter 1

SYSTEMS ENGINEERING FOR AAF COMMUNICATIONS

A broad study of the communications requirements for a proposed air warning system for European and Pacific theaters, including improvements to existing equipment, reduction of radio interference problems, development of new antennas, field-strength surveys, study of engine-generator noise, laying field wire by airplane, and many other subjects important to combat communications.

1.1 HISTORY OF THE PROJECT

Before the work under this project[a] was begun, the Bell Telephone Laboratories [BTL] had had experience in providing the Army Air Forces [AAF] with communications systems such as the extensive air warning and information center networks established and placed in operation in the United States before Pearl Harbor. When the Army Air Forces School of Applied Tactics was established in Orlando, Florida, BTL provided a complete information center communications system for training purposes.

When, therefore, plans were under way for operations overseas, the Air Forces needed the same kind of consultation. Now, however, the equipment would have to be portable and much of the service would have to be performed by means of radio instead of by wire telephone plant. The Signal Corps at that time had many communication components primarily suited to ground force operation, for the most part in the high-frequency (h-f) band and usually for continuous-wave (c-w) operation. The Air Forces on the other hand were primarily interested in very lightweight equipment that would be transportable by air and would be suitable for voice operation so that in the Air Warning Service either wire or radio telephone could be used at will, depending upon which was available at the time.

To implement a study of the systems aspects of communications, NDRC Project C-79 was set up, effective February 16, 1943. The objectives of this project were to "conduct research on engineering problems of the Army Air Forces communications systems and by oral or written communication cooperate in supplying information on special engineering problems to officers of the Army Air Forces."

The first assignment was improvement of the communications system (largely radio) for tactical Air Warning Service. This phase of the work was completed, and equipment for systems of the type recommended was procured by the Signal Corps for the Air Forces. The second assignment was similar to the first but concerned the complete tactical Air Force communications system. Work on this problem led to the conclusion that each tactical Air Force would have its own special requirements which would have to be considered individually. It was concluded that the best solution was to provide sufficiently complete and basic systems engineering information so that communication officers in the field could adapt and integrate into a satisfactory working system equipment available in the theaters.

A most valuable by-product of this effort was that it focused attention on the systems approach to communications problems and emphasized its importance to those responsible for Army communications. While the work on Project C-79 related solely to Air Forces' needs, the demonstration of the systems approach resulted in a subsequent direct request to the Bell Telephone Laboratories for the preparation of systems engineering information on a broader scale which, in turn, was instrumental in initiating a project between the Signal Corps and Bell Laboratories to provide this information in the form of Army manuals.[b]

Project C-79 covered a wide variety of subjects, including such problems as the avoidance of interference between radio sets in close proximity, the development of handy methods of estimating radio transmission over irregular paths, the use of h-f antennas adapted to transmitting short distances by means of sky waves, and the development of a new single-channel teletypewriter system which can be used on tactical radio sets working on a push-to-talk basis. The subjects of the individual reports making up the final report[1] indicate the wide range of problems considered under the project.

---

[a] Project C-79, Contract OEMsr-1018, Bell Telephone Laboratories, Inc., Western Electric Co., Inc.

[b] *Electrical Communications Systems Engineering* (TM11-486), preliminary issue dated February 25, 1944, revised edition, April 25, 1945. Much of the background for this comprehensive manual, particularly the v-h-f radio section, was related to the work done under Project C-79. The list of Signal Corps equipment was made into a separate manual, TM11-487, October 2, 1944.

[redacted]

1.2 PROJECT ACCOMPLISHMENTS

For tactical Air Warning Service, the h-f band was obviously overcrowded. Preliminary studies indicated that while both the 30- to 40-mc or 70- to 100-mc very-high-frequency (v-h-f) bands were suitable, commercial 30- to 40-mc equipment already available was the most readily adaptable. The principal contribution was in providing the Signal Corps and the manufacturer with information on means for minimizing interference effects.

The Air Forces were faced with the situation in which a large number of point-to-point radio circuits, each on a different frequency, converged at one location, resulting in a heavy concentration of transmitters and receivers. It became apparent early in this study that with current designs of both commercial and military v-h-f sets in the portable category, spurious transmitter radiations and receiver responses constituted a major interference problem when operating a number of sets in close proximity. The work dealt with the sources of these spurious effects, their magnitudes, set design factors, and practical aids such as the separation between sets and arrangements of antennas. The final step was the tedious process of determining workable groups of frequency assignments. This work provided a procedure for treatment of similar problems by both the Army and Navy.

In the military application of v-h-f radio sets, it is important to be able to predict whether or not a radio system can be expected to operate satisfactorily over specific paths. This problem was investigated, particularly for situations where the radio path includes intervening hills and so is not line-of-sight. Information for making such predictions was prepared in a form that does not require extensive technical knowledge for its understanding and is therefore very useful in the field. In this work, simple principles for the selection of good antenna sites were also stated.

In the latter phases of the study many of the same problems outlined above for the 30- to 40-mc band, were also studied for the 70- to 100-mc band for which equipment became available. Study of this equipment indicated that its characteristics with regard to mutual interference were comparable to those encountered in the lower frequency band. Furthermore, it did not possess certain operating and packaging features needed for the Air Forces' applications. A single-channel set more nearly meeting the requirements was designed.

Remote control arrangements were designed for both the 30- to 40-mc and 70- to 100-mc equipment discussed above.

The use of v-h-f, although well suited to the European theater, was not well adapted to situations frequently met in the Southwest Pacific where it was necessary to transmit through jungle territory or between islands beyond v-h-f range. An investigation was made which led to a recommendation that, for such cases, reliance be placed on transmission by sky waves in the 2- to 8-mc band, for which antenna arrangements which radiate well at high vertical angles were devised. Estimates were prepared of the field intensities necessary to override atmospheric static in various parts of the world.

A single-channel teletypewriter system was developed which permits changing a radio circuit back and forth at will between teletypewriter and voice operation, and which can be used on tactical radio sets working on a push-to-talk basis. Also included in the telegraph studies was the consideration of c-w Morse operation at v-h-f under conditions where speech is jammed.

Antennas for use with the various sets were considered. This included selection of suitable types, calibrations, and packaging arrangements. A somewhat related activity was the development of a lightweight 50-ft plywood antenna mast which can be erected by two men. This mast was procured in substantial numbers.

Another investigation was that of laying field wire from airplanes. This study, covering one method of paying out the wire, was carried on concurrently with work which the Bell Laboratories were doing under direct contract with the Army Air Forces covering other methods. Promising results were obtained from both these studies and the work was continued under a new contract between the Army Air Forces and the Bell Laboratories. The work under this contract, Project C-72, is summarized in Chapter 30.

Chapter 2

PROPAGATION STUDIES

2.1 EFFECT OF HILLS AND TREES ON RADIO PROPAGATION

2.1.1 Introduction

Under project C-79 a general study of the communications system of the air and ground forces was undertaken. Since the effect of obstructions as well as terrain characteristics is important from the standpoint of proper selection of radio-terminal sites, the investigations under Project 13.2-83[a] were essential as part of an overall study.

The final report[1] is divided into two major sections, Part I dealing with the transmitters, power supplies, methods of monitoring, field-intensity measuring equipment, antenna systems, calibration of the measuring equipment and radiation measurements. Part II is divided into sections dealing with the effect of trees, mountains, elevation of antenna system, and effect of receiver and transmitter locations.

2.1.2 Results of the Investigation

The woods used to study the effect of trees on radio propagation covered a sufficient area, and were such that the results should be valid for the great majority of cases in a region similar to that east of the Allegheny Mountains. The possible exception where these results would not apply would be a dense forest of tall trees with an abundant undergrowth, such as is found in some swampy areas and in tropical jungles.

2.1.3 Horizontal versus Vertical Polarization

Measurements at 28 and 116 mc with antenna heights of 19 to 29 ft demonstrated the general superiority of horizontally polarized waves in relatively broken country, both as to magnitude of the field intensity and uniformity in the presence of nearby interfering objects such as trees and houses. In the case of 116 mc, this is also true even in open country free from any outstanding obstruction.

A comparison of the results with theoretical computations for propagation over a smooth spherical earth shows that the horizontally polarized waves are not only attenuated less but also are in much closer agreement with theory for the antenna elevations used in the study. This is particularly true at 116 mc, where the actual field intensity for vertical polarization was 11 db lower than calculated while for horizontal polarization it was only 4 db lower. In rolling country vertically polarized waves were approximately 7 db less than horizontally polarized waves.

For horizontal polarization, trees of the type encountered would cause a loss in transmission of only about 10 per cent. For vertical polarization at 4 mc the loss is approximately proportional to the heights of the trees; in woods with 25-ft trees, approximately two and one-half times the power would be required, and with 50-ft trees ten times the power would be required to maintain communication at the same maximum range, as in open country. At 28 mc, the transmission loss due to trees was not serious. A lateral movement of antenna, a 5-ft increase in antenna height or a 100 per cent increase in power would overcome the effect of the trees. The signals were approximately normal if both antennas were above the foliage.

At 116 mc, vertical polarization is unsuitable in woods with 25-ft trees unless a position of minimum signal can be avoided. With 50-ft trees it would be necessary to operate from a position of maximum signal. A change in location of 2 ft would often cause the signal to change from maximum to minimum. On windy days the signal was often nullified even if a favorable location was employed.

2.1.4 Effect of Mountains and Hills

If a communication link is capable of maintaining communication over an unobstructed path for a distance so great that the field intensity of a wave propagated over an obstructing ridge is not less than the normal value, then the service range of the link will not be affected or impaired by the presence of the mountain. If this condition is not satisfied, a reduction in range will be experienced which at times may be very pronounced.

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[a] Project 13.2-83, Contract OEMsr-1010, Jansky & Bailey.

Furthermore, when extreme range is desired and a position on a ridge is not accessible, if the minimum distance requirement has been exceeded a skillful operator may pick an advantageous position behind a ridge and exceed the maximum distance obtainable from a position that avoids the ridge. This apparently anomalous behavior results from the signal gain due to constructive interference resulting from the increased number of ray paths arriving at the receiving point.

2.1.5 Effect of Antenna Height

Increasing the height of both transmitting and receiving antennas is equivalent to increasing transmitter power, *the governing factor being the proportional change in antenna heights,* unless the absolute transmitting heights are specified. For example, an increase in antenna heights from 19 ft to 29 ft is equivalent to a transmitting power increase of five times. Naturally, the same equivalent increase in transmitter power would not be realized if the antenna heights were increased 10 feet from a 100-ft elevation.

At 28 mc this gain is quite dependable under all circumstances. At 116 mc, under ideal conditions the results are consistent; in rolling terrain there is a positive gain but less than that experienced in level country; in woods in level country the results are erratic but on the average about what is found in level unobstructed country. Where the path of transmission crosses a mountain the gain is random and may be negative. If a fixed antenna height is to be used in mountains the average results will be as effective with antenna heights of 15 to 19 ft as they will be at greater heights. For maximum results a variable-height antenna is necessary, the gain being as great as 5 to 10 db above average and as much as 20 db above the minimum values with fixed-height antennas.

2.1.6 Choice of Transmitter and Receiver Locations

In wooded areas and over short distances (4 miles) there is little advantage in choosing a clearing rather than a location among the trees provided the foliage is at least 10 ft from the antenna. At greater distances a location in a clearing will give some advantage and a distance of 50 ft from trees 25 ft high is desirable.

The best location under all conditions is the crest of a mountain. In rolling country where a single predominant ridge is not available, the crest of a knoll is desirable. In some cases a site in front of the actual crest is better than on the crest, but the rear side should always be avoided. On a small level plateau there is little advantage in locating the site on the forward edge of the level area and no special effort is warranted in reaching it unless the situation is extreme.

2.2 RADIOTELEPHONE COMMUNICATION BETWEEN MOBILE UNITS

2.2.1 Introduction

The original purpose of Project C-30[a] was to determine the most desirable frequencies for radiotelephone communication between mobile vehicles over distances up to ten miles as well as to study the antennas in use on such vehicles and noise levels encountered by receivers under such conditions. The events of December 1941, however, required immediate military decisions, together with a tremendously accelerated procurement program and tended to shift the emphasis in the project away from the original purpose and toward the more practical problem of how best to use the available equipment and how to measure its performance.

A considerable portion of the final report[2] on this project is devoted to descriptions of equipment employed in the studies and methods of calibration utilized. In 1942 the Armed Forces were already using frequencies of 4 and 28 mc, approximately, for mobile ground communication, whereas the 116-mc frequency was mainly employed between plane and ground.

Throughout the project and in the final report[2] attempts were made to correlate the measurements and propagation data with theoretical curves because, by the use of such curves, the performance of radio systems over territory where studies have not been made can be predicted. Graphical methods of solving the complicated equations for ground-wave propagation over the surface of the earth were worked out and reported.

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[a] Project C-30, Contract OEMsr-174, Jansky & Bailey.

2.2.2 Work Accomplished

Following construction, procurement, and calibration of necessary transmitting and measuring equipment, field strength measurements on 4 and 28 mc were made in three areas, one in the Beltsville, Maryland, area characterized by rolling terrain, one in Bridgeville, Delaware, characterized as flat, and one near Twiggtown, Maryland, which was mountainous.

After these field intensity studies had been made, a means for measuring the free-space field from mobile antennas was worked out so that a study of the radiating properties of mobile Army antennas could be carried on with the object of comparing certain installations. The free-space field was useful as a figure of merit since it eliminated the effects of the ground and focused attention on the physical properties of the antennas.

The next step in the development of the project was a comprehensive study of the antenna characteristics of radio installations in Army vehicles including old- and new-style command cars, the M-3 light tank, and a 1½-ton panel truck with and without trailer. The effect of having the whip antennas vertical or in the semihorizontal running position was determined, showing that the radiating properties fell off badly with the antenna in the running position.

A study of ignition and atmospheric disturbances in mobile Army communications systems followed.

2.2.3 Results on 4 and 28 Megacycles

As a result of the studies made, it was recommended that transmitting installations be rated in terms of "twice the free-space field" produced at unit distance or the equivalent "unattenuated field at one mile." A method for doing so was worked out and is described in the final report.[2]

At 28 mc, for the receivers tested, it was found that input terminal voltages varied from 1.1 to 2.2 μv for each microvolt per meter of field intensity in which the antenna was immersed. The tests indicated that for vehicles in good condition, well isolated from sources of man-made noise, field intensities of the order of 1 μv per meter were entirely satisfactory for intelligible f-m communication; that in convoy service in the presence of vehicles with less satisfactory ignition-noise suppression, 4 to 6 μv per meter were necessary for an equivalent class of service. It was felt that situations in which more than 6 μv per meter would be required would be infrequent and could be eliminated by corrective measures on the vehicles or by a change in location. It was felt, furthermore, that atmospheric noise was not a factor in limiting communication over f-m circuits at 28 mc and could, therefore, be ignored.

When all factors were taken into account, the conclusion was reached that f-m signals in the 20- to 40-mc range were more consistent and reliable for ranges of 3 to 12 miles in level and rolling terrain than a-m transmission of equivalent power at 4 mc because they were not limited by atmospherics nor were they so susceptible to variations of ground conductivity. In mountainous terrain, the higher-frequency signals suffered in comparison with lower-frequency signals because of deep shadows and general reduction in the intensity of average signals.

Atmospheric noise was definitely the limiting factor in 4-mc communication.

2.3 STUDIES ON 116 MEGACYCLES

In general the work on 116 mc followed in outline that on 4 and 28 mc. Field studies were made in the rolling country near Beltsville, Maryland. As a result of these field measurements it was reported that reliable communication could be carried on over a distance of about 9 miles with a transmitting antenna not over 6 ft above ground, and with antenna power in the neighborhood of 25 watts, assuming that a 1-μv per meter received signal would give good communication.

Use of 116 mc was indicated only when the reduced size of the antenna structure or the secrecy secured by lack of sky wave would have tactical advantages and then only in gently rolling or reasonably flat terrain. Mountainous territory would affect the 116-mc signals to an extent even greater than that on 28 mc.

Chapter 3

FREQUENCY MODULATION VERSUS AMPLITUDE MODULATION

A comparison of the properties of a-m and f-m communication at very high frequencies, particularly for airborne use. Theoretical study of the maximum possible range of communication and of the properties of the two systems as to suppression of noise such as static and electrical disturbance arising from other equipment in the airplane, c-w jamming, etc.

3.1 PROPERTIES OF AMPLITUDE MODULATION SYSTEMS

In an a-m system, the final report[1] on Project 13-110[a] shows that, in the presence of random noise, the narrower the i-f pass band the better the performance, but that the impairment in performance due to widening the pass band is not great. For strong signals there is relatively little difference in the performance of a wide-band and a narrow pass band system. If the carrier strength is such that the carrier-to-noise ratio is unity, then the wide-band system has a ratio of signal to audible noise which is poorer by a factor of 2. Articulation tests made by the Psycho-Acoustic Laboratory, Harvard University, bear out this analytical criterion.

Random noise cannot be balanced out, for example, by using random noise from one channel to oppose and supposedly cancel the random noise in another channel. Against a man-made noise which is not random, balancing schemes may be of value.

If interference takes the form of short sharp pulses, their effect on intelligibility can be kept to a minimum, especially when they are limited or clipped in the receiver so that their maximum amplitude does not exceed the amplitude of the voice wave. Analysis shows that to keep the i-f transient pulse as narrow as possible, it is necessary that the i-f response curve be as broad as possible. Since, however, a flat-topped rectangular response curve is not realizable, the effect of the slope of the response curve is important. In fact, the slope of the sides of the i-f response curve is more important than the width in minimizing the width of the i-f transient.

If provision has been made for keeping the width of the i-f transient pulse as narrow as possible, the next step is to minimize its effect on the audio output. The final report[1] discusses clippers, counter-modulators, and balancers as being useful in reducing the effect on the audio output.

The analysis concludes that a good noise limiter and provision for minimizing the desensitization of the receiver due to a-v-c voltage developed by the pulses are essentials for a good communications receiver. The performance of an a-m receiver is not greatly affected by mistuning, provided the carrier and side bands do not come too close to the edges of the pass band of the receiver.

3.2 PROPERTIES OF FREQUENCY MODULATION SYSTEMS

Frequency modulation has certain noise-suppressing properties which are inherent. In the presence of random noise only, when the ratio of carrier to noise is large, the f-m system yields a signal-to-audible-noise ratio which is better than that of an a-m system by the expression

$$\frac{\sqrt{3} \times \text{i-f band width}}{B \times \text{a-f maximum frequency}},$$

where B is a factor depending upon the deviation ratio. For a deviation ratio of 5, B is 1.6; for a deviation ratio of 15, B is 1.3.

Wide-band frequency modulation is capable of producing an extremely noise-free audio output at the price of a high r-f signal strength. Narrow-band frequency modulation cannot attain so much freedom from noise but yields a usable output at a lower r-f signal strength and hence is better adapted to those applications where distance of communication is more important than perfection in the audio output.

The superiority of f-m over a-m systems is due to the fact that the noise voltage spectrum of the f-m system is triangular with zero value at zero frequency, increasing linearly with frequency. The result is that most of the noise output is of such high frequency that it is inaudible.

When the carrier amplitude is greater than the noise amplitude, the frequency deviation of the sum of the carrier and noise is the same as that of the carrier; but when the noise exceeds the carrier, the frequency deviation is the same as that of the noise alone. Thus noise effectively decreases the average frequency deviation and in this manner decreases the audio output. Analysis shows that, for weak carriers, noise suppresses the signal less when no limiter is employed. Thus there is an optimum level at which the limiter should operate.

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[a] Project 13-110, Problem No. 2, Contract OEMsr-1441, Harvard University.

So far as pulse noise is concerned, an f-m receiver equipped with a balanced discriminator is, under certain conditions, capable of complete suppression of pulses. Thus, there is no output when no carrier is present; if an unmodulated carrier exists at the center of the i-f band, no output is caused by a pulse at either peak of the carrier wave. These conclusions are true whether or not a limiter is used.

A pulse occurring at a position other than the peaks of the carrier tuned to the center of the i-f band produces noise from a balanced discriminator. Under this condition the noise suppression depends upon the limiter action and not upon the balance of the discriminator.

When the carrier is detuned from the center of the i-f band and there is no limiter, a pulse produces some output regardless of its position with respect to the carrier cycle. When the carrier is detuned in the presence of random noise, the a-f noise increases. Overmodulation, therefore, will increase noise, since the excessive frequency deviation producing the overmodulation is equivalent to excessive detuning.

3.3 COMPARISON OF F-M AND A-M SYSTEMS

In the presence of random noise only, it is certain that when the carrier is large compared with the noise, the f-m receiver is quieter in output. The minimum rms carrier strength at which this improvement is obtained is termed the threshold level and is about twice the rms value of the random noise. This threshold level is lower for a narrow-band system than for a wide-band system, and at carrier levels below the threshold of the narrow-band system the audio output of the narrow-band receiver is quieter than that of a wide-band set. For maximum range, therefore, the narrow-band system is preferred.

At carrier levels below the threshold, it seems that the f-m receiver is superior to an a-m receiver. Under this project, the receivers of SCR-508 f-m transmitter-receiver equipment were compared when the BC-603 receivers were employed first as an f-m receiver and then as an a-m receiver. The f-m performance was consistently better. At an r-f input of 1 μv, the ratio of signal-plus-noise to noise was 13 db for amplitude modulation and 37 db for frequency modulation.

Against pulse interference, the a-m receiver has no protection except that which may be added in the form of clippers or limiters. An f-m receiver with an i-f limiter is vastly superior in this respect.

With respect to jamming, there seems to be no marked superiority of amplitude or frequency modulation. Both can be jammed. So long as the desired signal is stronger than the jamming signal and does not take over the limiter, the signal from an f-m receiver is clean, but beyond this point interference rises rapidly. Both f-m and a-m sets are thoroughly jammed when the interfering signal is from 6 to 10 db greater than the desired signal. Amplitude-modulation receivers are less vulnerable to c-w jamming when the beat frequency is inaudible. Such interference can take over the limiter in the f-m set, but will take over the a-m set only when it is strong enough to desensitize the receiver through a-v-c action.

In the presence of pulse interference, it appears that the f-m receiver is quieter even if the a-m receiver has the best limiter possible.

Communication by f-m apparatus is more susceptible to multipath transmission when the difference in the time of transmission is of the order of the modulation cycle. When this time difference is of the order of one r-f cycle the effect is to alter the intensity of the modulated carrier identically in the two systems and not to distort the signal.

In the presence of a carrier in the f-m system, random a-f noise is reduced considerably and for this reason it is possible to balance a carrier voltage against a noise voltage so that a very weak carrier can open the "squelch" but noise cannot. On the other hand, in the absence of carrier a weak noise voltage can open the squelch. The balancing action is not possible in an a-m receiver.

With regard to size and weight there is definite advantage in frequency modulation.

3.4 GENERAL CONCLUSIONS

Narrow-band frequency modulation is preferable to either a-m systems or to wide-band systems for general communication purposes, especially if crystal control of the receiver is possible so that the r-f band width can be narrow. Frequency modulation requires somewhat closer tuning but since the improvement against noise is so great, precise tuning is not necessary. The capture effect prevents simultaneous reception of two signals and renders the f-m receiver somewhat less susceptible to jamming.

Noise limiters are necessary in a-m receivers. In an f-m receiver, on the other hand, the discriminator inherently cuts down noise and limiters may be employed in both the r-f and a-f circuits. At the time of this report, it had not been necessary to use limiters in the a-f end of the receiver.

An extensive bibliography is part of the final report,[1] which also contains numerous oscillographs, etc., giving visual comparison of the two systems of modulation.

Chapter 4

MICROWAVE COMMUNICATION SYSTEMS

A brief survey which evaluates microwave communication systems[a] in use, under development, or being considered in the United States prior to August 30, 1945. Pertinent facts concerning the established and contemplated systems for communication at carrier frequencies above 1,000 mc including a cross-band microwave system.

4.1 GENERAL CONSIDERATIONS

The usefulness of a given type of communication equipment is governed by a number of practical considerations, such as:

1. The type of communication service it supplies.
2. The economy with which it uses band space.
3. The intelligibility of reproduction and the noise level at the reproducer.
4. Its maximum reliable range.
5. Its ease of operation.
6. Its size, weight, cost.
7. Its ruggedness and ease of maintenance and repair.

These practical aspects depend upon a number of technical considerations which will be reviewed in turn.

4.1.1 Types of Operations

A microwave system may embody one or more of the following types of operation:

General call (every station in a network receives a call addressed to one of them).

Selective call (only the called station receives a call).

Unidirectional operation (flow of intelligence in only one direction in a given communication circuit).

Duplex operation (flow of intelligence in both directions simultaneously in a given communication circuit).

Relay operation (instantaneous retransmission of an incoming message).

Simplex operation (only one wave of intelligence on a given r-f carrier).

Multiplex operation (several waves of intelligence on a single r-f carrier).

4.1.2 Types of Emission

The type of emission is an important consideration in microwave transmission. The emission may have any one of a number of forms, such as:

Continuous wave, amplitude modulated [a-m].

Continuous wave, frequency modulated [f-m].

Continuous wave, phase modulated [p-m].

Pulsed wave, pulse-amplitude modulated [p-a-m], in which equally spaced short pulses undergo variations in amplitude corresponding to the instantaneous value of the modulating signal voltage.

Pulsed wave, pulse-length modulated [p-l-m], in which pulses of constant amplitude are varied in length (i.e., duration) in accordance with the instantaneous value of the modulating signal voltage.

Pulsed wave, pulse-number modulated [p-n-m], in which short pulses of equal amplitude and equal spacing are transmitted. Signaling is accomplished by emitting a varying number of these pulses in accordance with the instantaneous value of the modulating signal voltage.

Pulsed wave, pulse-frequency modulated [p-f-m], in which short pulses of equal amplitude are transmitted at a variable rate corresponding to the instantaneous value of the modulating signal voltage.

Pulsed wave, pulse-position (phase) modulated [p-p-m], sometimes referred to as *pulse-time modulated* [p-t-m]. Pulses of constant amplitude and length are advanced and retarded in time from a uniform repetition rate by amounts of time which are proportional to the instantaneous value of the modulating signal voltage. The displacements of the pulses are measured either in time or in terms of the displacement of the r-f wave in radians.

4.1.3 Spectra and Spectrum Widths

The spectrum width W of an a-m signal is twice as great as the highest frequency of the modulating signal f_m if the modulation does not exceed 100 per cent. Thus,

$$W = 2f \qquad \text{(for a-m)}. \tag{1}$$

The width of the significant spectrum of an f-m signal is, to a first approximation, slightly more than twice the maximum frequency deviation Δf (the maximum displacement of the frequency when voice signal is loudest). Thus,

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[a] Project 13-110, Problem 9, Contract No. OEMsr-1440, Harvard University.

$$W \doteq 2\Delta f \quad \text{(for f-m)}. \qquad (2)$$

A reasonable approximation for the spectrum of a pulsed wave can be determined from the pulse duration a, without regard to the modulation. Assuming the pulses to be rectangular, the width W of the central portion of this spectrum is

$$W \doteq \frac{2}{a} \quad \text{(for pulsed emission)}. \qquad (3)$$

In a-m transmission, for $f_m = 3$ kc, $W = 6$ kc.

In f-m transmission in the microwave region the most suitable value for Δf, and therefore the value of W, will be influenced by a number of factors. The most suitable value had not yet been determined at the time of this report. Values of Δf of the order of 100 kc are contemplated.

The values of a employed in the existing pulsed-emission equipments result in calculated central spectrum widths W of the order of 3 mc, with a tail on each side of the central part having appreciable height over a range of the order of 3 mc. The actual spectra are apt to be broader and perhaps unsymmetrical as a result of undesired reflections due to imperfect impedance matches and from undesired alteration of the carrier frequency due to the pulsing of the transmitter. It appears that simplex pulse transmission could be characterized by a considerably narrower spectrum than that just described, although the spectrum would still be many times broader than that of a-m or f-m transmission.

This may be reasoned as follows. In simplex pulse transmission it is not suitable for the pulse repetition rate f_p to be less than two or three times f_m, which sets a lower limit on f_p. For example, let us assume a minimum value of f_p of 8 kc. A duty factor must next be assumed. A large duty factor tends to narrow the spectrum and at the same time to lessen any increase in transmission range resulting from the use of pulses, and vice versa. If a duty factor of 5 per cent is assumed, the pulse length a is 6.25 μsec, and, by equation (3), $W \doteq 0.32$ mc.

4.1.4 Multiplexing of Signals

The various schemes for multiplexing of signals on a single r-f carrier may be divided into two general groups: frequency-division multiplexing, and time-division multiplexing.

In *frequency-division* multiplexing, several intelligence signals are combined into one signal which is applied to the transmitter as a single modulating wave. Usually the frequencies of the signals to be combined are elevated to different degrees and are then added together. Two examples follow:

"Spiral-four" audio-frequency multiplexing. Four voice-frequency signals, each ranging from 200 to 2,800 cycles per second, are combined into a modulating signal having a frequency range from 200 to 11,600 cycles per second. Channel 1 contains an ordinary voice spectrum. Channels 2, 3, and 4 contain single side-band (lower side-band) suppressed-carrier signals.

Supersonic-frequency multiplexing. In this scheme the modulated subcarrier frequencies are supersonic frequencies; for example, 30, 40, and 50 kc.

Time-division multiplexing is best described by means of an example. In the AN/TRC-5 equipment, the emission consists of a 2.0-μsec marker pulse followed by eight 0.4-μsec channel pulses which, in the absence of modulation, are equally spaced in time. The complete cycle of nine pulses requires 100 μsec. Once each 100 μsec, each of eight voice-frequency circuits is "sampled," and the channel pulse assigned to it is position-modulated.

Frequency-division multiplexing has been used only in connection with continuous-wave emission, although it is applicable to all types of emission. Time-division multiplexing is especially suited to pulsed transmission.

In general, multiplexing provides "trunk-line" service and some degree of privacy. It results in little if any economy of band space, other factors being equal. Calculations[1] indicate that to retain a fixed value of signal-to-noise ratio, the theoretical output spectrum must be widened directly as the number of signals combined by multiplexing. In the case of frequency-division multiplexed a-m transmission consisting of one carrier and an upper and lower side band for each voice channel, carriers are eliminated but guard space must be provided between the frequency ranges allocated to the various voice channels. In f-m transmission, calculations show that the value of Δf and hence that of W must be increased in the same proportion as f_m to maintain a fixed signal-to-noise ratio. In time-division multiplexing of pulsed transmission, the minimum number of pulses required for simplex transmission must be multiplied slightly more than N

times if N voice channels are to be accommodated. Hence, multiplexing of pulsed transmission requires a decrease in a and an increase in W.

CROSSTALK PROBLEM

An inherent difficulty in connection with frequency-division multiplexed transmission is crosstalk in the form of modulation products resulting from nonlinearity in the amplitude characteristics of the equipment. In a-m transmission this is due to nonlinearity in the tube characteristics, and in f-m transmission to nonlinearity in the phase response of the circuits. In time-division multiplexed transmission under proper conditions only one voice signal is handled at a time, so that crosstalk ideally is impossible. However, distortion due to tube or circuit characteristics is not impossible in pulsed transmission, for it may occur in the tubes and circuits of the modulating portion of the transmitter.

4.1.5 Establishment of Carrier Frequency

In microwave communication, the carrier frequency must be readily and precisely selectable and extremely stable. A precision of the order of one part in 10^6 or better is required to realize fully the possibilities in the microwave region predicted by theory.

Crystal control yields the required stabilization, but a huge number of crystals would be required for flexible operation of a system consisting of hundreds of stations and hundreds of frequency channels if each channel is to be available to every station. Crystal saving has been attempted in connection with several v-h-f systems; for example, a few crystals may be used to provide a large number of frequencies through the processes of beating and multiplication. Another possibility is to mix a crystal-stabilized signal with the output of a tunable oscillator and use the sum or difference frequency as the carrier frequency. In this system all but a very small part of the carrier frequency is crystal-controlled.

If crystals are used, there still remains the problem of selection of the correct frequency multiple. At the high level of multiplication required, this calls for considerable equipment. Also, if the r-f output power tube is not a power amplifier, crystal stabilization may be had only in the form of an automatic-frequency-control (a-f-c) circuit, which decreases the stability obtained.

4.1.6 Use of Band Space

While in theory the number of channels possible in a given amount of band space depends solely upon the type of emission and the desired signal-to-noise ratio, in practice the number of possible channels depends also upon other practical factors. Perhaps the most important are the accuracy with which the transmitter carrier frequency can be reset and stabilized, and the stability of the local oscillators in the receiver. The following data are based upon the premise that all such practical difficulties are surmountable.

Table 1 represents an attempt to determine approximately the relative ultimate possibilities of the use of band space in simplex communication. Guard bands of estimated minimum practicable size have been allowed between the spectra. These values represent combinations of calculations and estimates, which is as far as one can go with safety at the present stages of both theory and practice.

TABLE 1. Estimates of the minimum carrier-frequency spacing for simplex transmission in the microwave region, assuming the absence of all practical difficulties.

Type of emission	Description of transmission	Estimated minimum space between carriers
AM	Carrier and two side bands; $f_m = 3$ kc; 4-kc guard space allowed between spectra.	10 kc
FM	Carrier and two side bands; $f_m = 3$ kc; $\Delta f = 3$ kc; 4-kc guard space between spectra.	10 kc
Pulsed	$f_m = 3$ kc; $f_p = 8$ kc; duty factor $= 5\%$; $a = 6.25$ μsec.; $W \doteq 0.32$ mc; 0.18-mc guard space allowed between spectra.	0.5 mc

The data in Table 2 show the estimated minimum amounts of band space required per voice channel in various forms of multiplex transmission, again assuming the absence of all practical difficulties.

It is emphasized that the values shown in the two tables must be regarded only as estimated limits. Also, the values do not mean too much on a relative basis because the amounts of difficulty encountered in trying to achieve these limits will not necessarily be in the same ratios.

It is possible theoretically to operate several pulsed transmitters on the same carrier frequency simultaneously without interference by using different pulse repetition rates. However, this provides little if any possibility of increase in the number of voice channels in a given amount of band space because the transmitter which operates at the highest repetition rate would call for a channel width comparable to the sum of the channel spaces that would be required if each transmitter were operated at the same minimum repetition rate with nonoverlapping spectra.

The shift in carrier frequency caused by the doppler effect is investigated in an appendix of the final report.[7] The frequency shift is not likely to be great enough to affect the values in Tables 1 and 2, for it is not probable that the pass band of the receiver will be narrow enough for a noticeable effect on communication to be produced even in the most extreme case, where a frequency shift of the order of 12 kc is noted.

Table 2. Estimates of the use of band space in multiplex transmission in the microwave region, assuming the absence of all practical difficulties.

Type of emission	Description of transmission	Required band space per voice channel
AM	4 voice channels, frequency-division multiplexed by use of the spiral-four scheme; $f_m = 11.6$ kc; 6.8-kc guard space allowed between r-f spectra; carriers 30 kc apart.	7.5 kc
FM	4 voice channels, frequency-division multiplexed by use of the spiral-four scheme; $f_m = 11.6$ kc; $\Delta f = 11.6$ kc; 6.8-kc guard space allowed between r-f spectra; carriers 30 kc apart.	7.5 kc
Pulsed	8 voice channels, time-division multiplexed; f_m of each voice channel = 3 kc; one marker pulse for each 8 channel pulses gives $f_p = 9 \times 8{,}000$ or 72 kc; average duty factor = 5%; $a \doteq 0.7$ μsec, $W = 2.8$ mc; 1.2-mc guard space allowed between spectra; carriers 4.0 mc apart.	0.5 mc

4.1.7

Required R-F Power and Its Development

A maximum range of 100 miles is generally considered satisfactory for microwave communication. A distance of 25 miles frequently is regarded as sufficient.

The amount of r-f output power required is a subject on which there is some theoretical disagreement, and is a matter that is governed largely by the kind of antenna permitted by the type of communication service desired. If a directional antenna may be employed at the transmitter or the receiver or at both, an average r-f output power of the order of 1 or 2 watts may be sufficient to achieve the desired range. Experience indicates that if omnidirectional antennas are employed at both stations, the average r-f output power should be of the order of 50 watts or more to achieve maximum range.

At present the most practical types of power tube available for microwave communication are the klystron and the magnetron. The klystron is limited to an average output power of the order of 2 watts. High-power magnetrons intended for pulsed transmission have been well developed. Tunable magnetrons intended for c-w transmission, having an output power of the order of 1 kw or more, have recently been designed. Certain existing models of klystron and magnetron lend themselves readily to frequency modulation.

4.1.8

Cross-Band Signaling

The cross-band principle of signaling has been used in carrier systems and may be used in any of the radio communication systems to be described.

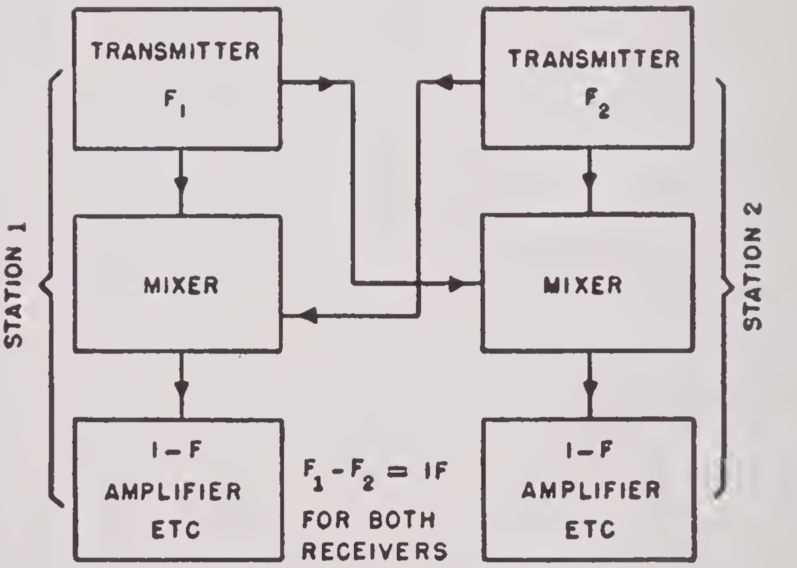

Figure 1. Cross-band system of communication.

Cross-band signaling will be described with the aid of Figure 1. Stations 1 and 2 are alike in all respects, except that the antenna system at either station may be directional or not, as required. Transmission and reception go on simultaneously at each station. If a beamed antenna is used for transmission, either the same antenna or an identical antenna aimed in the same direction is used for reception. The two stations are in contact when they are within range and when their carrier frequencies F_1 and F_2 differ by an amount equal to the intermediate frequency employed by both stations. At each station, a portion of the modulated output signal is applied to the mixer instead of a signal from a local oscillator for frequency conversion. The signaling system is operationally bilateral and offers the following features of operation:

1. Establishment of contact becomes reasonably certain to the calling party when he finds that he can hear himself talk. The presence of the other station's carrier makes his own voice audible, and this implies that his own signal is also being received by the other party. A request for verification of contact usually is unnecessary.

2. Duplex operation, i.e., two-way communication similar to an ordinary telephone conversation, is achieved. Push-to-talk operation is eliminated; either party can interrupt the other at will.

3. The strength and quality of the voice signals heard by the two parties are practically the same, so that each one usually knows how well he is being received without asking for a report.

An outstanding virtue of cross-band signaling in the microwave region lies in the fact that it eliminates the necessity for having a separate first local oscillator for mixing in reception and the attendant necessity of stabilizing its frequency. An associated advantage is that if either party's oscillator is slightly off frequency, this weakens the signal equally at both stations, and readjustment of the frequency of either oscillator improves or weakens the reception at both stations.

4.1.9 Protection from Interference and Jamming

Obvious measures that may be taken in the microwave region to forestall interference and jamming of communication service are (a) the use of high power, (b) the use of directive antennas, and (c) the use of widely separated frequencies for transmitting and receiving. Other measures are (d) the use of very narrow pulses and hence the widest practical spectrum, (e) adjustable signal-amplitude selection circuits (clippers and limiters) to aid in the elimination of pulsed interference, and (f) some form of manual or automatic volume control which can be used to prevent overloading of amplifiers by a c-w signal.

4.2 SUMMARY

4.2.1 Comparisons of Microwave Communication Equipment

At the time of this report, there has been little production or use of microwave communication equipment.

The few experimental mobile-ground, shipborne, or airborne microwave communication sets which have been built conform basically to the existing push-to-talk v-h-f equipment, and their characteristics are in such an indefinite state that critical comparisons are not timely.

No broadly applicable model of microwave communication equipment has been developed. The sets so far developed are relatively bulky equipment intended primarily for transportable point-to-point radio relay service and are not suitable for other applications.

It appears that there would be use for a simple, small, lightweight, rugged, versatile, inexpensive microwave communication set that could be produced easily in large quantities. The delay in the development of such a broadly applicable set has not been occasioned by lack of need or desire, but because the development of radar equipment was given higher priority. The radar developments have led to the creation of numerous microwave components which now could be reassembled into a system designed for communication.

4.2.2 The Choice of the Type of Emission

It appears that a set intended for broad application and for production and use in large quantities should employ c-w emission. Continuous-wave emission permits simplicity in equipment and achieves economy in band space. At this time, it is easier to produce an f-m signal than an a-m signal, but this condition may disappear. There are some indications that f-m may remain preferable to a-m. For example, f-m may offer more promise in regard to signal-to-noise ratio and antijamming properties.

The use of pulsed emission for relay and trunk-line service appears to be satisfactory. The ultimate possibilities in multiplexed c-w transmission apparently have not been determined. It is possible that developments in equipment, for example, in the field of carrier-frequency stabilization, may make multiplexed c-w transmission more practical than it appears to be at present. It would then be a competitor with multiplexed pulse transmission because of the concomitant economy of band space and simplicity of equipment.

Theory indicates that p-p-m is the best all-around choice among the five types of pulsed emission that were considered. Also, extensive laboratory tests made by the Federal Telephone & Radio Corp. [FT&R] and the Radio Corporation of America [RCA] indicate

[redacted]

that when maximum protective devices are used, the antijamming effectiveness of p-p-m and p-f-m approach equality. According to an FT&R report, the p-p-m system has many advantages over the p-f-m system. They may be listed as follows.

1. No change of average power during transmission.
2. Greater degree of privacy.
3. Possibilities of multiplexing, which is extremely difficult to obtain with p-f-m.
4. Ease of center pulse-frequency stabilization.
5. Ease of obtaining and applying blocking potentials.
6. Further advanced state of development of the p-p-m system.

4.2.3 Stabilization of Carrier Frequency

In pulsed transmission, the stability and accuracy of the carrier frequency need not be so great as in c-w transmission because of the breadth of the spectrum. When the production of an extremely accurate and stable carrier frequency becomes a simple matter, this advantage of pulsed emission over c-w emission will disappear, and also the possibility of several thousand c-w channels in a given microwave frequency band will become a reality.

While a crystal-control scheme may be applied to microwave communication equipment, there is much promise in cavity stabilization using the microwave frequency discriminator described by R. V. Pound.[6] This a-f-c system is simple and straightforward, is not limited to steps of frequency, and will make available at every station an unusually large number of stabilized carrier frequencies. The success of the microwave discriminator a-f-c system will depend largely upon the development of a tuned cavity whose resonant frequency may be very finely adjusted in a resetable manner, and whose frequency is affected negligibly by changes in temperature.

4.2.4 Use of Cross-Band Operation

Cross-band operation requires simpler and less equipment. The first local oscillator of each receiver is eliminated, and the a-f-c system that stabilizes the transmitted carrier frequency also stabilizes the signal used for the first conversion in reception.

Cross-band operation could be instituted in connection with almost any one of the existing or contemplated equipments. There would be no great advantage, however, in applying it to pulsed or to multiplexed point-to-point trunk-line systems. It would be particularly useful in connection with a communication system composed of hundreds or thousands of stations, any one of which is to be able to contact any other station. Cross-band operation is achieved readily now that r-f power-dividing waveguide sections such as the "magic T" have been developed.

From the operator's point of view, cross-band operation simplifies and speeds up the establishment of contact. The called party does not have to touch or adjust anything when called, but can simply start talking. With any other existing or planned two-way radio system having the same individual-channel privacy offered by the cross-band system, the called party has to make a frequency selection of some sort before he can start responding, and he has no indication that he is being heard by the calling party until the calling party acknowledges his response. The cross-band system would minimize interference between stations and would result in economical use of the available band space.

PART II

3,000-MC COMMUNICATION

Chapter 5

FIELD TESTS AND 3,000-MC EQUIPMENT

A study of the suitability and possibilities of 3,000-mc communication over land and over sea water; development of an omnidirectional and a directional communications system.

5.1 STATE OF THE ART

When this work was started in June 1941, there had been little if any mobile communication on microwaves. Point-to-point circuits had been operated on wavelengths in the centimeter region, in particular the 1,600-mc (18-cm) link across the English Channel; microwaves had been extensively used in radar. The object of this research,[a] therefore, was to investigate the possibilities of the microwave region for mobile communication.

5.2 INTRODUCTION

As a preliminary to the development of actual 3,000-mc communication systems, a study was made of propagation characteristics and circuit requirements, with the following results:[1]

1. Vertically polarized waves were found best over salt water; over land there was little difference between vertical and horizontal polarization.
2. At these frequencies it was found that a substantially optical line-of-sight path was required, that signals were greatly absorbed by trees and houses, and that the signals disappeared very quickly beyond the optical horizon.
3. Power gains of 100 to 200 are obtainable with antennas of relatively small dimensions, making it practicable to communicate over distances of 15 to 20 miles with power of the order of 1 watt. This high gain results in high directivity, which makes unauthorized interception difficult.
4. Where high directivity cannot be used, the transmission power must be increased considerably. (A 30-watt omnidirectional system developed under Project C-42[2] had about the same communication performance as a 400-mw directional system.)
5. Two-way voice communication was demonstrated by means of equipment used for the preliminary tests (Project C-24) over a distance of 14.5 miles.

5.3 PROPAGATION OVER SALT WATER

To study propagation at 3,000 mc over salt water, a klystron transmitter with a power output of 2 watts and a 30-in. parabolic reflector with a power gain of 23 db over a half-wave dipole were installed on land, and a receiver consisting of a crystal converter with a 1221-Y local oscillator and a 30-mc i-f amplifier were installed on a 38-ft Coast Guard boat. The receiving antenna consisted of a small horn with a 1-ft square opening having a gain of 15 db over a half-wave dipole.

In the first test the transmitter antenna was installed on a hill south of Port Jefferson Harbor, 192 ft above sea level, and directed toward Bridgeport, Connecticut. Difficulty was experienced in keeping the boat in the beam of the transmitter, as some of the runs were made after dark and because no markers were available for checking the position. In a second test, the antenna was moved to a bluff on Mt. Misery, Long Island, almost directly over the water and 125 ft above it. The course ran east from there to Horton's Point. Figure 1 shows the profile of this course; signal strengths using vertical polarization over the course are given in Figure 2. On this run voice communication could have been carried out to grazing incidence at 24 miles. Keyed tone modulation could have been carried out somewhat further, but no signal was heard at a distance of 34 miles.

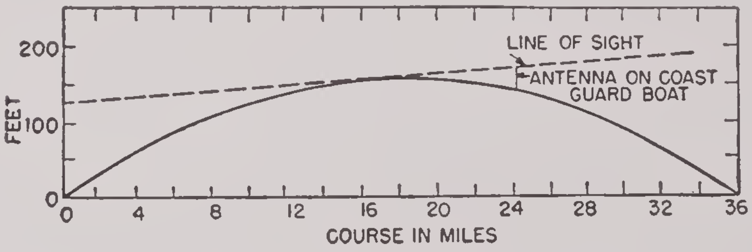

Figure 1. Profile, Mt. Misery Point and Long Island Sound based on 4/3 of earth's radius.

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[a] Projects C-24 and C-42, Contract Nos. OEMsr-32 and OEMsr-442, RCA Laboratories; Project C-2, Contract No. NDCrc-75, RCA Manufacturing Co., Inc.; Project C-10, Contract No. NDCrc-191, Westinghouse Electric & Manufacturing Co.

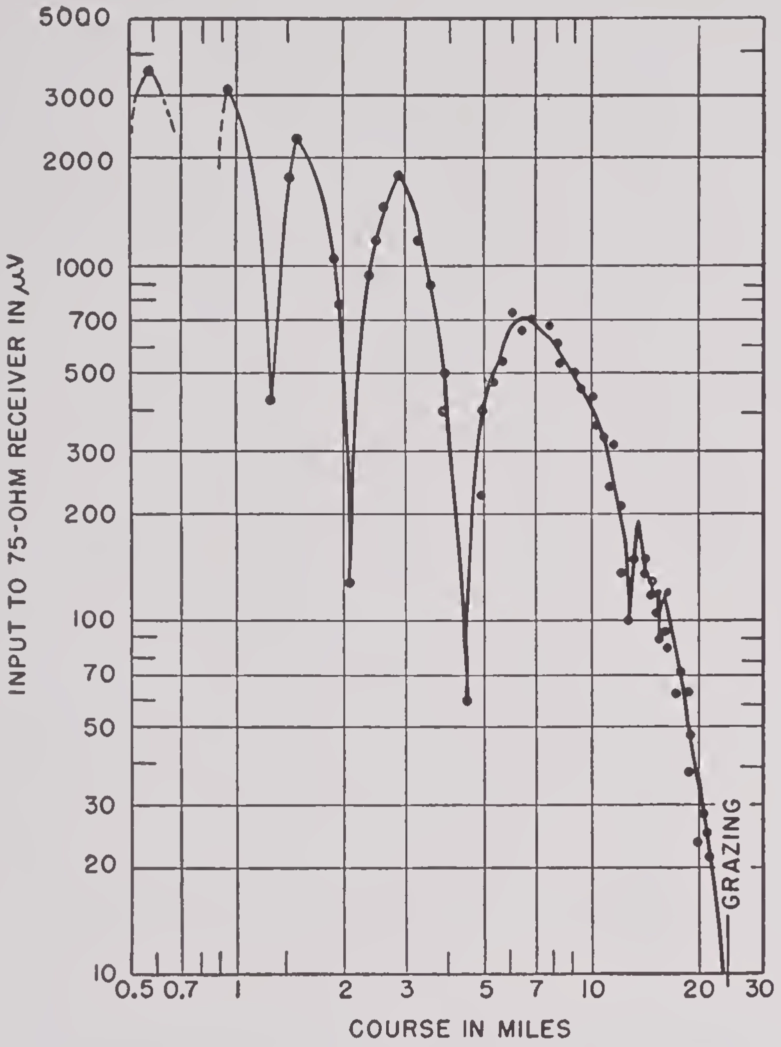

Figure 2. Signal strengths with vertical polarization over sea water; transmitter at Mt. Misery, Long Island, course east to Horton's Point.

A run using horizontal polarization was made over approximately the same course (to Mattituck, Long Island), the resulting measurements (Figure 3) indicating the maximum range to be about the same as for vertical polarization.

These tests of the two types of polarization showed that with horizontal polarization the minima were very deep and short, the signal being inaudible at these points indicating almost complete cancellation between the direct and the reflected rays. This means that the coefficient of reflection for water for this polarization is almost 100 per cent for the rather small angles which the rays made with the water. With vertical polarization the minima were not nearly so deep.

The conclusion to be reached from these tests is that, for a communication circuit over salt water, vertical polarization is definitely indicated.

5.4 PROPAGATION OVER LAND

Three tests were made to investigate propagation over land. These consisted of a variable antenna height test, a test of signal strength as received in an automobile at various locations, and recordings of signal strength between fixed locations versus time.

Variable Antenna Height Test

The klystron transmitter and reflector were mounted in a small elevator suspended from one of the towers located at Rocky Point. The receiving antenna used was a horn with an opening of 2.14x2.65 ft and having a gain of 23 db over a half-wave ($\lambda/2$) dipole. This was mounted 70 ft above ground and connected to the receiver by means of a waveguide. The transmitter and receiver were 14.5 miles apart. The gain of the parabolic receiving antenna was also 23 db over a half-wave dipole. As the 2-watt transmitter and antenna were raised and lowered, signal strengths were measured at the receiver. The maximum received signal occurred within a few feet of the height calculated from the profile. Vertical polarization was employed. (See Figure 4.)

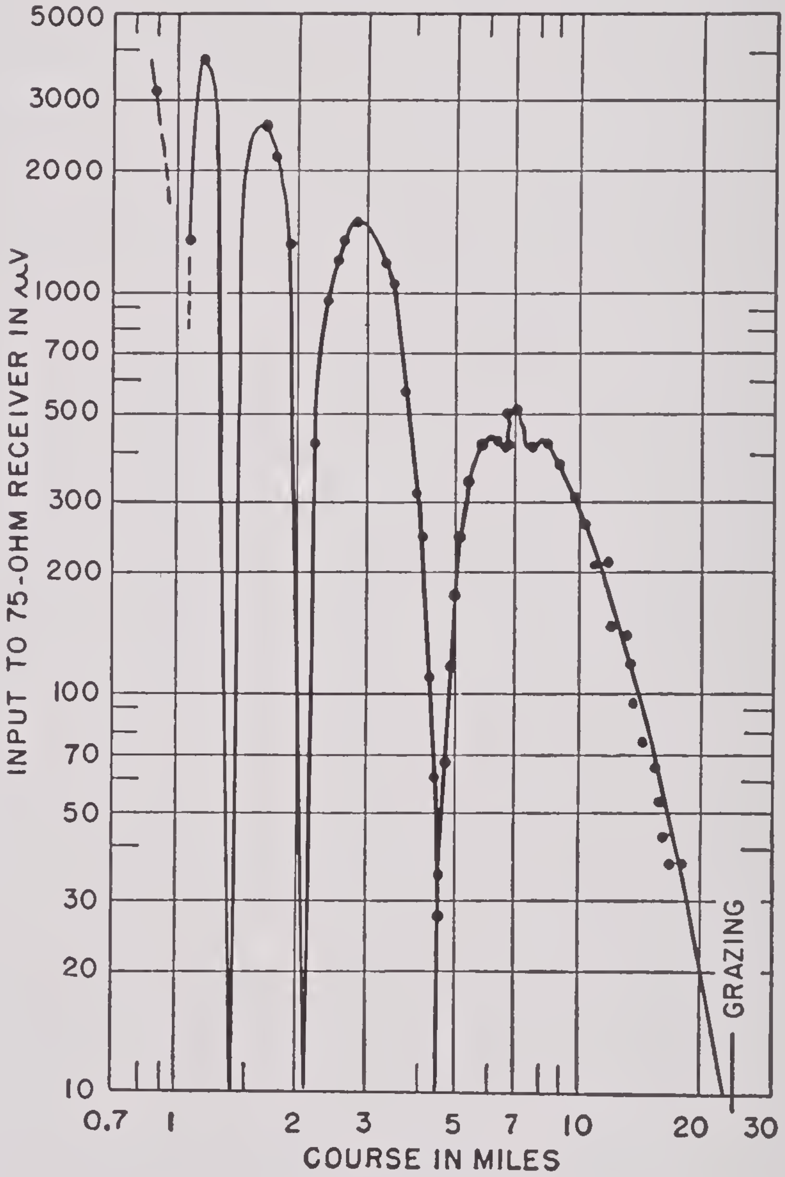

Figure 3. Received signal strengths for horizontal polarization; transmitter at Mt. Misery, course east to Mattituck, Long Island.

Automobile Tests

The transmitter antenna was mounted 80 ft above ground (Building No. 10 at Rocky Point) and was fed through a waveguide. The receiver with a small horn antenna was transported by automobile to various locations. The signal strengths were quite variable, in fact "shotgun" in pattern. The signal disappeared quite often when the car was moved a foot or two. Local reflections and absorption were quite noticeable from objects in the vicinity.

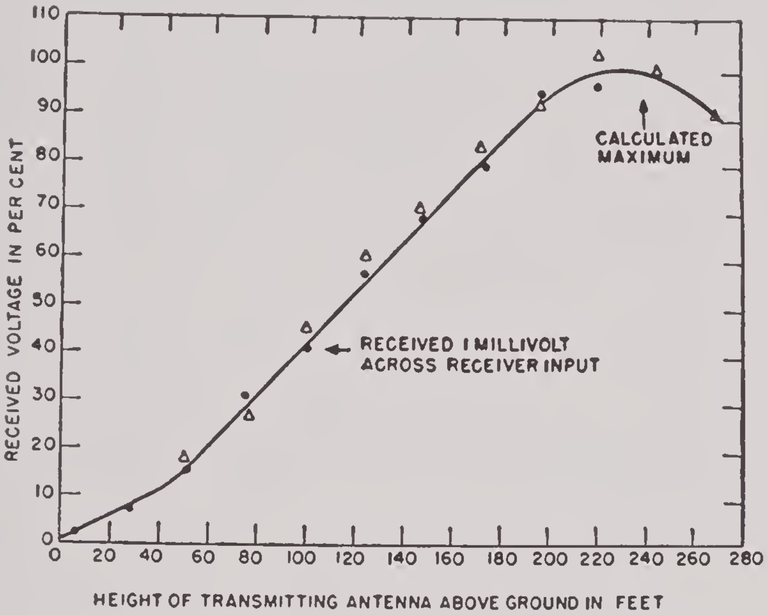

Figure 4. Signal strength versus height for 10-cm signals, vertical polarization, Rocky Point to Riverhead, Long Island.

Recording Tests

Two sets of recordings of the signal strengths versus time were made over a distance of 14.5 miles (between Building 10 at Rocky Point and Building 9 at Riverhead). The receiver employed a 23-db horn antenna. The recordings were of chief interest because of the severe refraction occurring on the evening of November 5. (See Figure 5.) Similar conditions were observed over nearly the same path on 500 mc which was also being recorded.

Severe refractions were noted again on the night of February 15 and 16. No effect was noticed due to heavy rain, fog, or light snow. No selective fading or static was noticed at any time. Ignition noise entered through the 30-mc i-f amplifier but this could be removed with adequate shielding.

The conclusion reached is that 10-cm communication over land would be satisfactory with either horizontal or vertical polarization, provided the terminals of the system were within line-of-sight of each other.

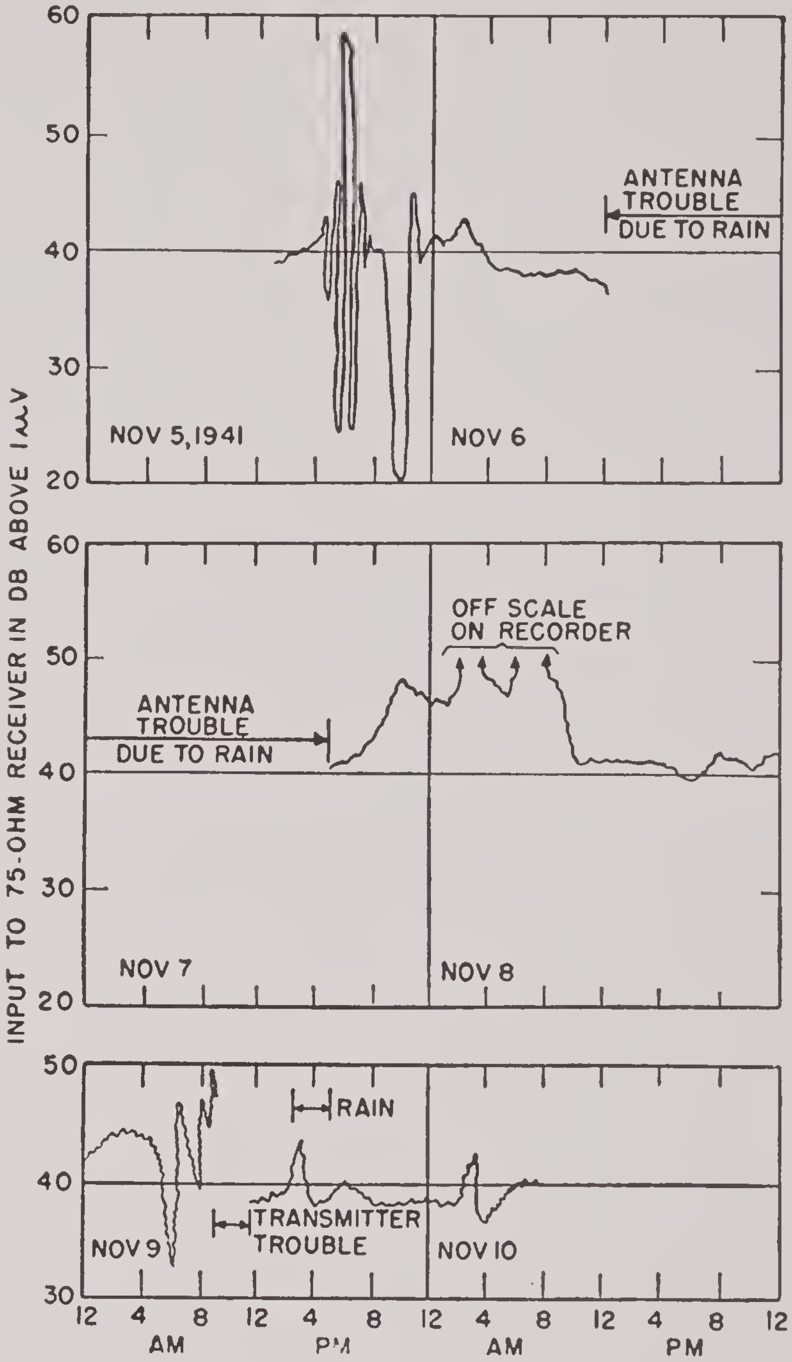

Figure 5. Recording of 10-cm propagation, vertical polarization, Rocky Point to Riverhead, showing abnormal conditions on evenings of November 5 and 9.

5.5 CRYSTAL-RECEIVER TESTS

Since it has been suggested quite often that a simple crystal detector followed by an audio-frequency amplifier could be used for portable reception, some tests were made to determine the feasibility of such a simple receiver. The tests showed that such a receiver had a very poor *equivalent noise side-band input* [ENSI][b] in the neighborhood of 100 μv across 75-ohm input terminals, which is over 40 db poorer than the crystal heterodyne receiver employed in the tests or 65 db poorer than an ideal receiver in which the noise would be proportional to the absolute temperature T and the frequency band Δf.

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[b] ENSI (equivalent noise side-band input) is the equivalent input magnitude of all random noise which is transferred to the output circuit, and therefore, of all such noise within the frequency side-bands passed by the receiver.

This receiver with a 15-db antenna was carried about in the vicinity of an omnidirectional transmitter having an antenna with a gain of about 6 db with an input of 2 watts. Over a distance of about one mile, reception was very unsatisfactory due to the presence of trees and other obstructions. Behind large trees no signals could be heard, and even in fairly clear places the standing-wave pattern was such that the receiver had to be moved to keep away from zero-signal points.

The results indicated that reception with such a simple receiver could be expected up to a distance of about a mile but that continuous reception by a walking or moving person was not practical.

5.6 EQUIPMENT EMPLOYED

Two transmitters and three receivers plus accessory equipment were employed in these tests and a considerable number of measurements were made on stability, modulation capabilities, receiver sensitivity, and frequency modulation versus amplitude modulation.

5.7 TRANSMITTERS

One of the transmitters furnished for the preliminary investigation of 3,000-mc propagation characteristics was a Westinghouse klystron unit supplied under Project C-10.[3] The other was a Western Electric unit which is described in Chapter 6.

Table 1 gives salient facts about the 2 transmitters.

TABLE 1. Comparison of 3,000-mc transmitters.

	Western Electric transmitter	Westinghouse klystron transmitter
Size	$66\frac{1}{2}$x22x17 in.	37x21x15 in.
60-cycle power input	775 watts	450 watts
Power output at 3,059 mc	4 watts	4 watts
Plate supply ripple	—60 db	—68 db
Plate supply bounce	About 3.5 volts of 1,500 volts	About 6 volts of 2,500 volts
Accuracy of frequency setting	±1.0 mc	No provision

The frequency drift of the two transmitters during the warm-up period is shown in Figure 6. Both transmitters were operated from a voltage-regulated power supply. The instantaneous frequency stability of the two transmitters was comparable, the klystron transmitter being frequency-modulated ±0.2 mc when it was amplitude-modulated 35 per cent, the Western Electric transmitter having a band of 1 mc. These figures were obtained in listening tests measuring the width of the frequency spectrum with a receiver. The wider band of the Western Electric transmitter can easily be explained by the harmonics in its 16-kc pulses (CFVD).[c]

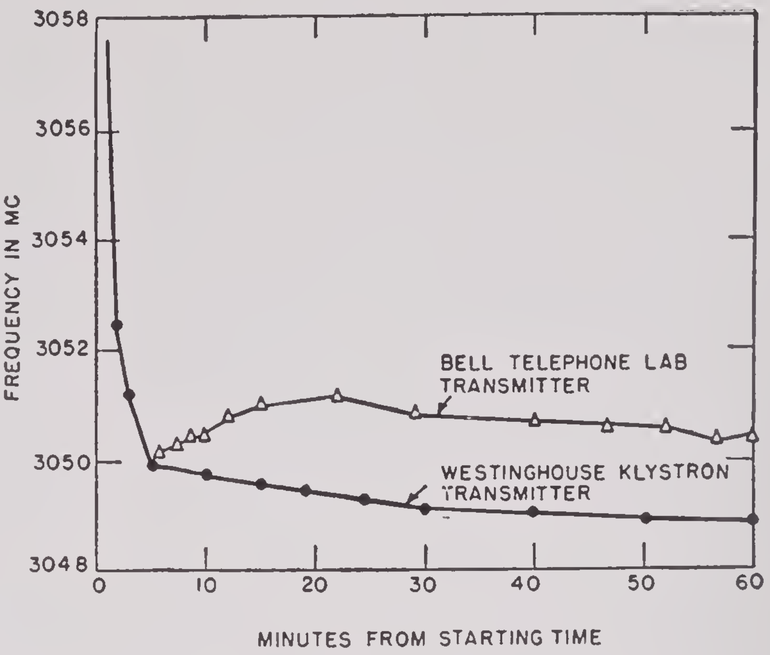

FIGURE 6. Frequency drift after starting 3,000-mc transmitter.

5.7.1 Frequency versus Temperature

Frequency and ambient temperature measurements on the klystron transmitter showed that there was no definite correlation between these two factors, indicating that other causes existed for the frequency variations. At the time of the measurements, varying weather conditions disturbed the impedance match between the transmitting antenna and its waveguide feed which, in turn, presented a variable load to the transmitter, at times causing the transmitter to stop oscillating.

When the horn antenna was used, greater stability was experienced, probably because it was more waterproof than the parabolic reflector antenna. With the horn, recordings indicated that the transmitter stayed within the 2-mc pass band of the recording receiver over a temperature variation of approximately 63 to

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[c] See Chapter 6. CFVD indicates a continuous-frequency variable-duration pulse system.

75 F. During this period the transmitter did not stop oscillating of its own accord. Both transmitter and receiver were operated from a voltage-regulated power source.

An increase of 1 per cent in the 115-volt power supply to the klystron transmitter caused —180 parts per million change in output frequency. This corresponds to a 540-kc frequency shift at 3,000 mc.

5.7.2 Automatic Frequency Control

The commutator of the automatic frequency control of the Western Electric transmitter was replaced with the balanced detector circuit shown in Figure 7. In this circuit a synchronous motor drives a four-pole variable condenser at 1,800 rpm. Thus, when the frequency is off the resonant frequency of the monitor cavity, a 120-cycle voltage is generated and is impressed in push-pull upon the grids of the balanced detector. Plate voltage for the detector is obtained from the input to the magnetic field rectifier supply and while it should be a 120-cycle sine wave, the 60-cycle full-wave rectified voltage was satisfactory. The performance of the automatic frequency control system was the same as with the commutator and had no sliding contacts. It used the same number of tubes as the original circuit.

5.7.3 Transmitter Modulation

A modulator similar to the Western Electric transmitter modulator was built for the klystron transmitter. In this system the transmitter is keyed with 50 per cent square dots under the no-modulation condition. At the peak of the audio cycle, the keying mark goes to 100 per cent (up) and to zero (down). An ordinary a-m receiver with a diode detector followed by a 5-kc low-pass filter will receive such transmission.

The advantages of such a CFVD system are:

1. Frequency modulation of the transmitter is minimized which greatly eliminates distortion in the receiver due to its not having a perfectly flat band-pass characteristic.
2. Full modulation capabilities are easily realized without undue distortion.
3. Loading and tuning of the transmitter are less critical.

In this system the pulse frequency was 21 kc; the build-up time of the pulses was 1.0 μsec. The klystron grid was driven with a pulse amplitude of 100 volts peak to peak. Since there was no direct current on the grid, the effective modulating voltage was half this value because the klystron is cut off on the negative swing. By applying positive bias voltage to the klystron, the maximum positive voltage reached by the grid is increased with a corresponding increase in average power. Adding 15 volts positive to the grid, in this case, increased the modulating peak voltage to 65.

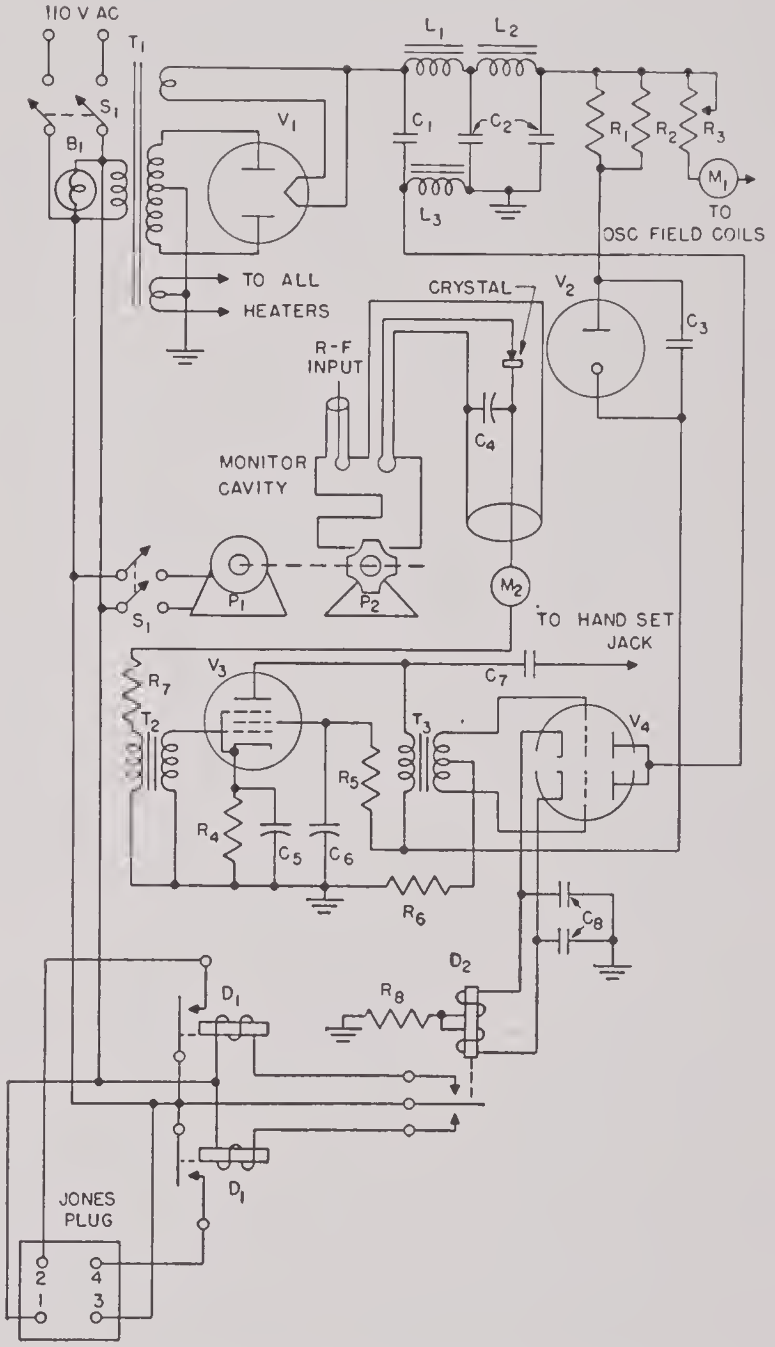

FIGURE 7. Automatic frequency control for Bell Laboratories transmitter.

5.7.4 FM versus AM

To compare frequency modulation with pulse modulation, the frequency of the klystron was modulated by varying the plate voltage through the output transformer of the modulator placed in series with the plate supply to the klystron cathode. The resultant amplitude modulation in the receiver was limited out. A modulating voltage of 50 volts peak to peak gave a peak frequency deviation of 300 kc. This corresponded to 100 per cent modulation with a receiver pass band of 600 kc.

A signal-to-noise ratio improvement of $60\sqrt{3}$ should be obtained using frequency modulation; the narrow frequency band-pass of the receiver, however, made it apparent that automatic frequency control would be necessary to keep the transmitter from drifting out of the pass band of the receiver.

5.7.5 Push-to-Talk Circuit

To turn the transmitter off, the length of the coaxial line to the antenna change-over switch was adjusted so that when the switch was opened the reactive load placed upon the transmitter caused it to stop oscillating. This maintained constant input power to the transmitter cavity. The transmitter did not come on at the same frequency at which it went off, however, and special means had to be taken to increase the power to the transmitter during the stand-by period. This was accomplished by increasing the plate voltage from 1,600 to 1,775 volts by means of a relay operated by the change-over switch.

5.8 RECEIVERS

Three receivers were tested during this study. One of them employed a crystal converter with a 1221-Y oscillator; another used a special tube containing a resonant-cavity oscillator into which was coupled a hairpin circuit tuned to the signal frequency, the output being amplified in an 8-mc amplifier; the third receiver used a beam deflection tube.

5.8.1 Crystal Converter

This receiver (Figure 8) was made from parts supplied by the Western Electric Company. Small changes in oscillator frequency were made by means of a screw driver on one of the oscillator plate inductors. The 30-mc i-f amplifier (IR-202), developed and manufactured by RCA under Project C-2,[5] had a mid-band frequency of 30 mc, an equivalent band width of 2.75 mc at full gain and 3.25 mc at somewhat lower gain. The automatic gain control characteristic was quite flat above 15 to 20 μv input.

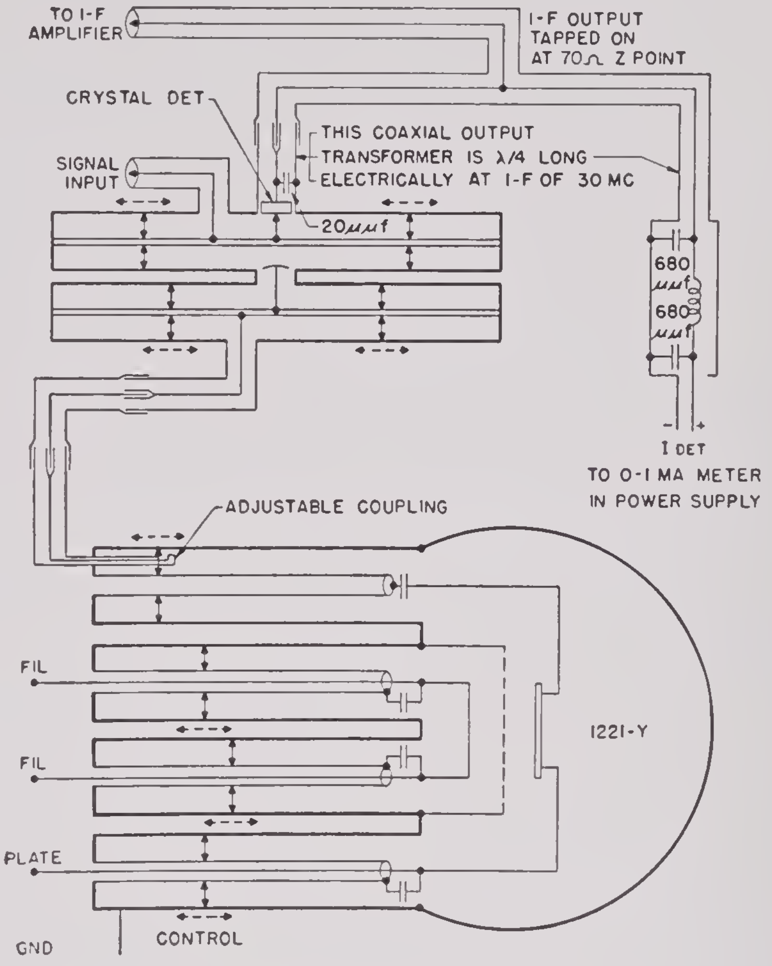
TO I-F AMPLIFIER
I-F OUTPUT TAPPED ON AT 70Ω Z POINT
CRYSTAL DET
THIS COAXIAL OUTPUT TRANSFORMER IS λ/4 LONG ELECTRICALLY AT I-F OF 30 MC
SIGNAL INPUT
20μμf
680 μμf
680 μμf
I DET
TO 0-1 MA METER IN POWER SUPPLY
ADJUSTABLE COUPLING
FIL
FIL
1221-Y
PLATE
GND
CONTROL

Figure 8. Ultra-high-frequency converter using 1221-Y oscillator and crystal detector.

5.8.2 Receiver Performance

Frequency Stability. The frequency change was 2,070 kc or 28 parts per million per degree (ppm/°C) over an ambient temperature variation from 36 to 81 F. A +1 per cent change in plate voltage caused a change of +55 parts per million. A +1 per cent change in filament current caused a frequency change of +73 parts per million; changing the line voltage to the oscillator power supply by +1 per cent produced a frequency change of +69 parts per million.

Input Noise Characteristic. The ENSI of this receiver with its i-f amplifier was 0.8 μv across 75 ohms for an r-f band width of 10 kc. This is 23.2 db poorer than the best ideal receiver ($KT\Delta f$ base). This measurement was made with a 3,060-mc r-f carrier well above the peak noise level in the i-f amplifier, so that the intermediate-frequency threshold was not a factor in determining the noise. The ENSI of the i-f amplifier itself was 0.21 μv under the same conditions, showing that most of the noise came from the crystal converter.

Sensitivity. To determine the weakest signal that could be received, taking the threshold into account, an a-f amplifier having an equivalent band width of 5 kc was used. Under these conditions a 100 per cent modulated signal of 3.7 μv across the 75-ohm input gave a signal-to-noise ratio of 10 db.

RCD-23 RECEIVER

The 3,000-mc converter of this receiver made by the RCA Manufacturing Company was a resonant cavity oscillator plus an i-f amplifier with a mid-band frequency of 8 mc. Due to the fixed oscillator cavity, tuning over a limited range was accomplished by varying the cavity voltage and the input tuning.

Stability. Oscillator stability with temperature was less than 7 ppm/°C. Changes in cavity and cathode voltages of +1 per cent produced frequency changes of +400 parts and —100 parts respectively.

ENSI. At 3,060 mc, ENSI was 3.8 μv across 75 ohms for an r-f band width of 10 kc using a signal above the peak noise of the i-f amplifier. This is 36.8 db poorer than the ideal receiver. With a signal of 25 μv across the input (below the i-f threshold) the ENSI was 5 μv for a 10-kc band.

BEAM DEFLECTION CONVERTER RECEIVER CR-301

In this superheterodyne receiver the frequency converter was a beam deflection tube (H-2214-2A or H-2243-1) providing separate deflection circuits for the signal and for the local oscillator injection. The signal circuit cavity was toroidal in shape and was so constructed that the signal voltage between opposite faces of the cavity deflected the electron stream as it passed through a slot in the cavity. The local oscillator voltage was applied to a pair of rods which deflected the electron stream before it passed through the signal cavity.

Between the signal cavity and the collector plate was a slot bisected lengthwise by a fine target wire. When the electron beam was at rest (without oscillator or signal deflection voltages) and properly focused and centered, most of the electrons were removed by the wire. Deflection of the electron beam to either side of this central partially masked position produced an increase in collector current. Thus when the beam was deflected through this central position the slope of the curve of collector current versus deflection voltage reversed. This reversal of slope produced a large value of conversion conductance. The collector plate was treated to accentuate secondary emission, thereby providing one stage of electron multiplication.

The local oscillator of this receiver was a ZP-446 equipped with a regulated power supply.

The ENSI of this receiver was 0.9 μv across 75 ohms for a 10-kc band width. This is 24 db poorer than an ideal receiver.

5.9 WAVEGUIDES AND ANTENNAS

Several types of waveguides were studied with a view to finding the most practical method of feeding antennas at a distance from the transmitter. In the region studied (3,000 mc), feed lines of about 100 ft of ordinary coaxial cable have too high a loss. Horn and reflector types of antennas were examined as to their practicability for communication systems at these frequencies.

5.9.1 Waveguides

Because of the relative ease of bending rectangular guides and the better control of polarization compared to circular guides, two types of rectangular guides were used:

1. Copper rain spouting $3\frac{1}{16}$x$2\frac{3}{8}$ in. with 0.020-in. walls, corrugated lengthwise, was used in 10-ft lengths. The measured loss was 0.39 db per 100 ft at $\lambda = 9.8$ cm. This figure compares with the calculated loss of 0.36 db.
2. Commercial bronze rectangular tubing 3x1½ in. with 0.064-in. walls, hard-drawn, 90 per cent copper, 10 per cent zinc. Calculated loss was 0.85 db per 100 ft at $\lambda = 9.8$ cm; measured loss was 0.79 db.

Various bends were tried. If the radius of curvature of the inner surface was 6 in. or more, negligible reflection occurred. Tapered sections of various lengths showed reflections varying around 1 per cent.

Small amounts of water in the waveguides increased the losses greatly.

5.9.2 Antennas

Two horn antennas with an aperture of 2.65x2.14 ft were made of plywood and lined with copper foil. They had a power gain of 23 db over a $\lambda/2$ dipole at 3,000 mc. Another horn with a 1-ft square aperture having a calculated gain of about 15 db was used on both the land and the salt water surveys.

The paraboloid reflectors 30 in. in diameter were used for transmission. One was fed with a $\lambda/2$ vertical dipole; the other reflector had the same feed system but with the addition of a parasitic tuned dipole in front of the radiator substantially to eliminate direct forward radiation. The directive pattern of this antenna is shown in Figure 9 where it will be seen that the pattern is quite narrow. It measures ±7° wide 6 db down. The horn antennas were down 6 db at ±8° in the horizontal plane and at ±6½° in the vertical plane.

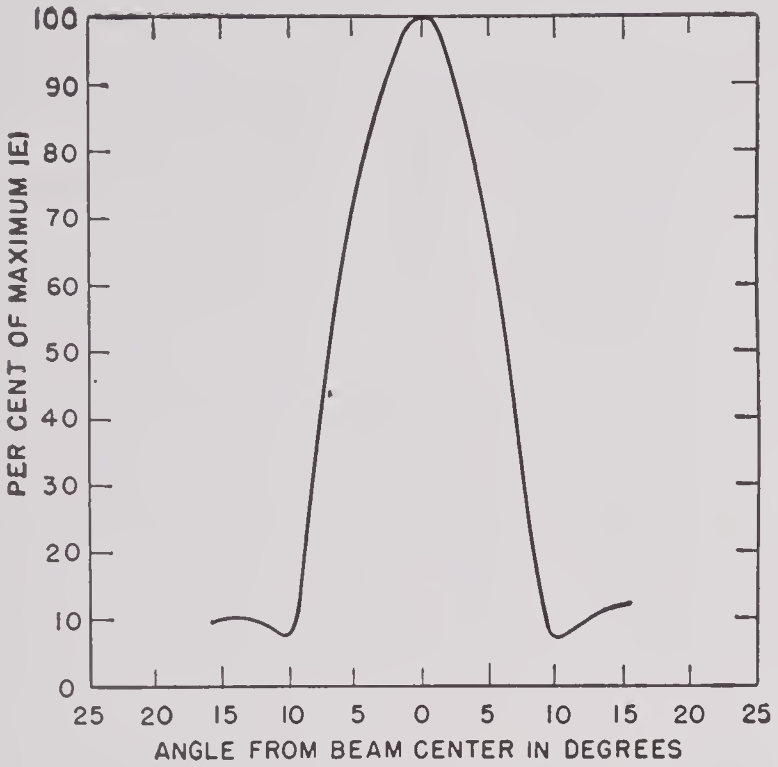

Figure 9. Radiation pattern of 30-in. reflector with dipole and parasitic unit.

5.9.3 Loss in Plywood

In testing antennas it was desirable to locate the units inside some sort of building and for this reason it was important to know the effects of plywood on radiation from antennas. Tests were made, therefore, to find out something about the losses in plywood.

The effects of the material upon a free space wave would be very difficult to measure experimentally. Such properties, however, may be indirectly determined by placing slabs of the material in a rectangular waveguide.

Various thicknesses of plywood were obtained by first making a thick slab from ⅛- and ¼-in. pieces of board held together by small bakelite dowels. By using various combinations of thicknesses, steps in thickness of approximately ⅛ in. were obtained.

5.9.4 Measurement Procedure

A rectangular waveguide was terminated in a horn at its far end. Beyond the horn was a device for measuring radiated power. A standing wave detector was near the transmitter. The guide was perfectly matched by means of a variable capacitance. Then the slabs were inserted between the standing wave detector and the horn. The power output and both the magnitude and position of the standing waves back of the plywood were measured.

From these data the ratio of power input to power output and the magnitude of the coefficient of reflection could be determined. From the magnitude and position of the standing wave, the impedance looking toward the load from the back side of the plywood slab was calculated. These data are plotted in Figures 10 and 11. Note that the impedance is plotted in the complex plane and forms a spiral. The efficiency of transmission for a ½-in. (12-mm) slab of dry plywood is about 60 per cent. Soaking the slab over night in water reduced its efficiency to about 43 per cent.

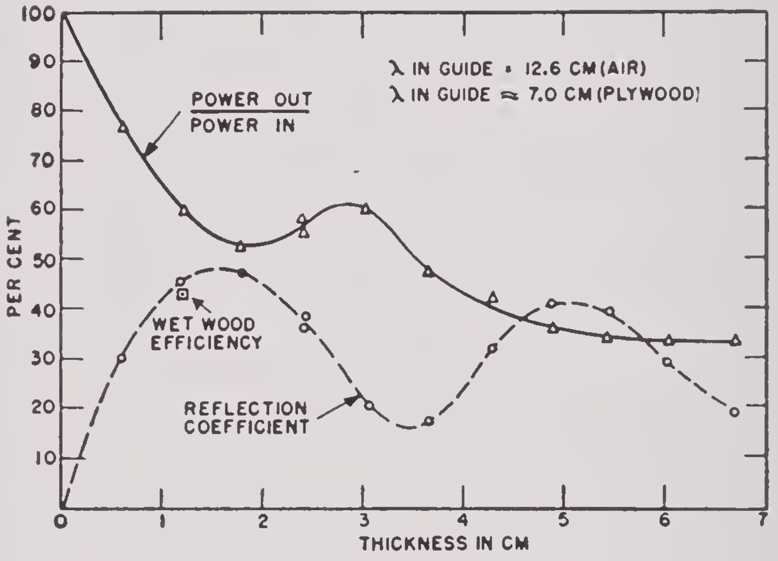

Figure 10. Transmission efficiency and reflection coefficient, dry plywood at 9.7-cm wavelength.

Minimum reflection takes place when the thickness of the wood is approximately λ/2 for a wave propagated through plywood in the 2x3-in. waveguide. If the material were lossless the reflection would be zero for the λ/2 thickness.

From these data the dielectric constant of the plywood is estimated at about 2.33. There is some uncertainty about this figure since the data were not consistent, probably because this wood is nonhomogeneous and the properties of pieces probably vary substantially. Another method,[6] used at the Massachusetts Institute of Technology [MIT], in which standing waves in a waveguide terminated in a conducting sheet are measured, indicates a value of 1.9 for the dielectric constant of plywood at λ = 6 cm. It is quite possible, however, that the plywood tested at MIT was drier than the samples examined under this project.

5.9.5 Power Measurement

Power measurements were made with a load resistor consisting of 41 in. of No. 38 Nichrome wire (41.6 ohms per ft) wound in a 20-turns-per-in. square thread on a $1\frac{1}{16}$-in. diameter brass cylinder. The square thread was 0.027 in. deep by 0.020 in. wide. The wire was kept in the center of the groove by a 0.01-in. silk thread. A brass sleeve ($\frac{1}{16}$-in. wall) was pushed over the cylinder. This sleeve had a coaxial fitting at one end with the inner conductor connected to one end of the Nichrome wire. The other end of the wire was connected to a terminal on the cylinder. The cylinder had a well to receive a thermometer.

The section of the waveguide to which the resistor was attached was connected to the waveguide from a 3,000-mc transmitter. The waveguide-to-coaxial-line transformer was adjusted so that the load resistor was matched to the waveguide (approximately 50 ohms). The temperature rise with time was compared with the curves of Figure 12 to determine the power into the resistor. If the power flowing through the waveguide was known, the crystal probe could be calibrated.

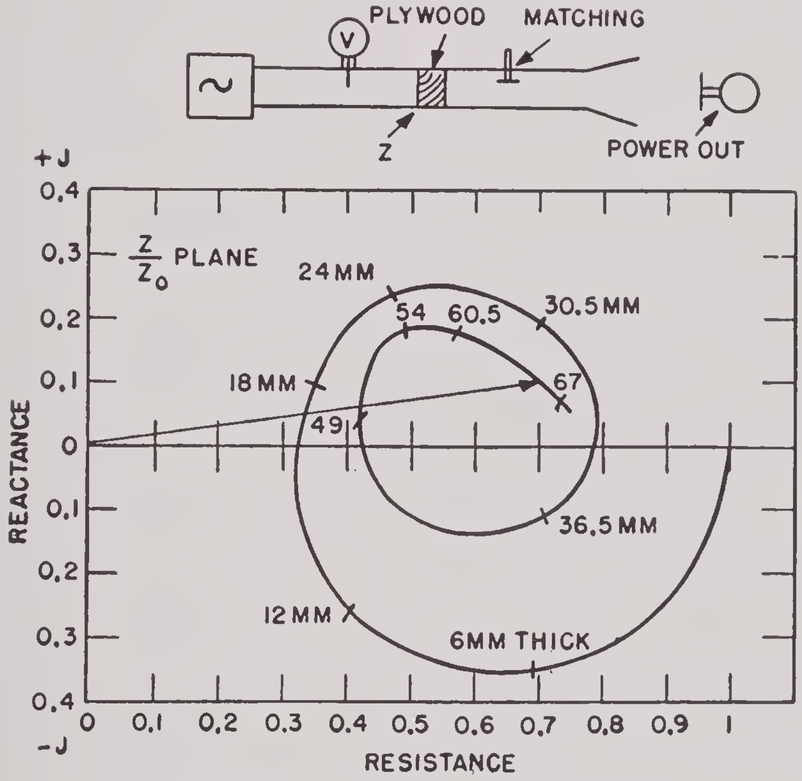

FIGURE 11. Impedance versus thickness, plywood in waveguide.

5.10 OMNIDIRECTIONAL MICROWAVE TELEPHONE

Following the preliminary investigations carried out under Project C-24 and described above, Project C-42 was set up to develop an omnidirectional telephone system and a directional microwave telephone, using the data secured under C-24.

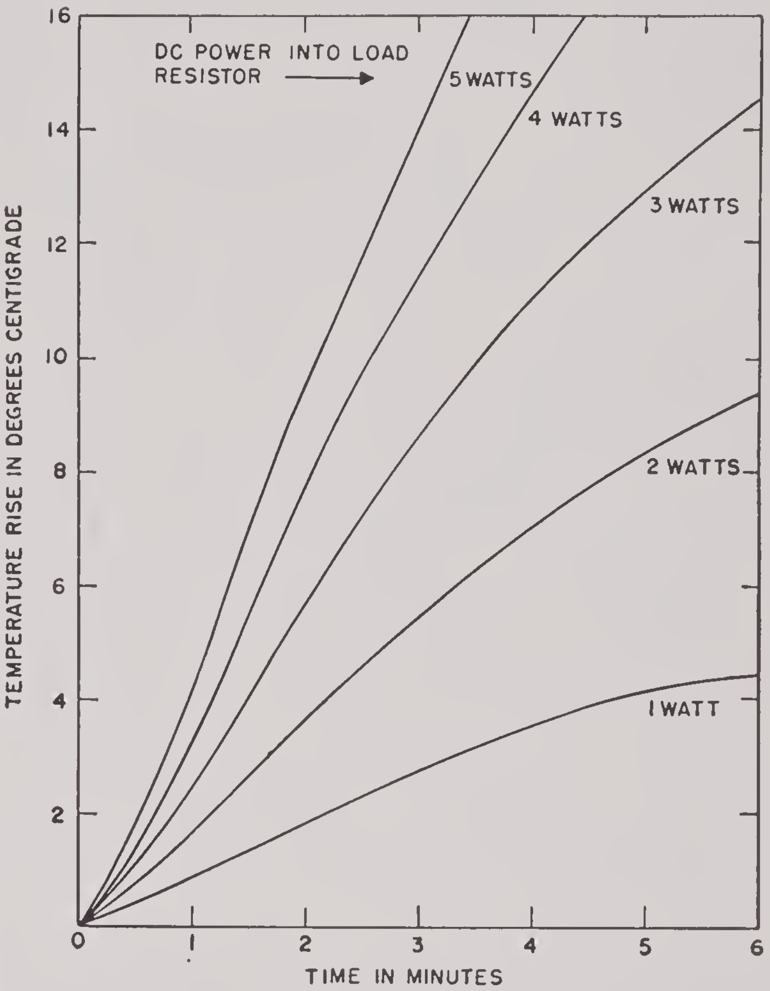

FIGURE 12. Calibration of ultra-high-frequency load resistor used in power measurements.

The omnidirectional system was desired by the Navy as an additional telephone communication channel having a certain amount of secrecy, operating somewhere in the region of 1,200 to 2,700 mc and having a range in any direction of 10 miles, all stations transmitting on the same frequency. The equipment, designed to operate at 1,400 mc, consists of two transmitter-receiver combinations mounted with their associated control equipment. Overall system tests were made between Rocky Point and Riverhead, a distance of 14.5 miles. The transmitter antenna, a vertical doublet, was mounted 77 ft above ground. The receiver antenna was mounted 49 ft above the ground and consisted of a vertical doublet with a parabolic reflector. With 30-watt average power output from the transmitter at Rocky Point, the power delivered to the receiver at Riverhead was 3.3 $\mu\mu$w. A block diagram of the system is shown in Figure 13.

5.10.1 Transmitter

The transmitter tube was a magnetron having an average output of 30 watts on a carrier frequency of

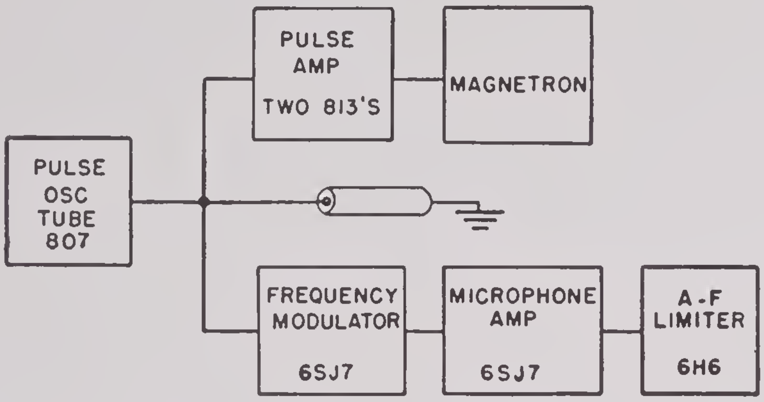

Figure 13. Block diagram of nondirectional system working on 1,400 mc.

about 1,400 mc. The output signal consisted of 0.8-μsec pulses having an unmodulated pulse rate of 20 kc. The pulse rate was frequency modulated, the maximum deviation being ±3 kc. The frequency-modulated pulse system provided an improved signal-to-noise ratio at the receiver over that obtained by other pulse systems and simplified the transmitter design. Furthermore this system lowered the noise threshold by 12 db compared to a c-w system of the same power, representing an increased effective range for the system. (A circuit diagram of the pulse transmitter is given in Figure 14.)

The output from a single-button carbon microphone (Western Electric Type F3) was sufficient to modulate the transmitter 100 per cent. An audio limiter prevented overmodulation due to high input from the microphone but permitted a reasonable per cent modulation for weak microphone input voltages.

The transmitter operated from 110-volt 60-cycle mains and required 750 watts at a lagging power factor of 0.9.

5.10.2 Magnetron Details

The r-f oscillator and output tube was an air-cooled multianode magnetron, the output carrier frequency

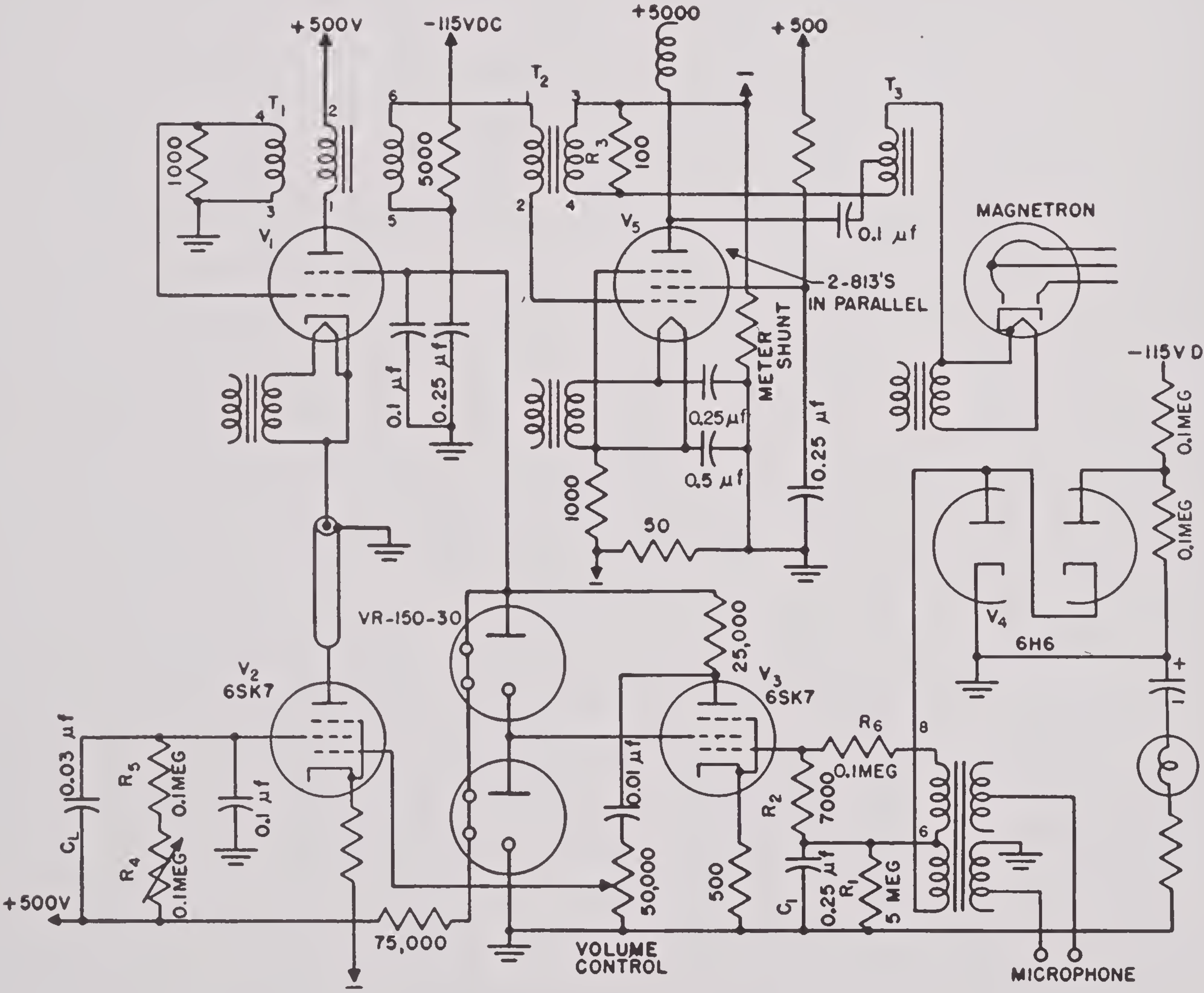

Figure 14. Circuit diagram of 1,400-mc pulse transmitter for nondirectional system.

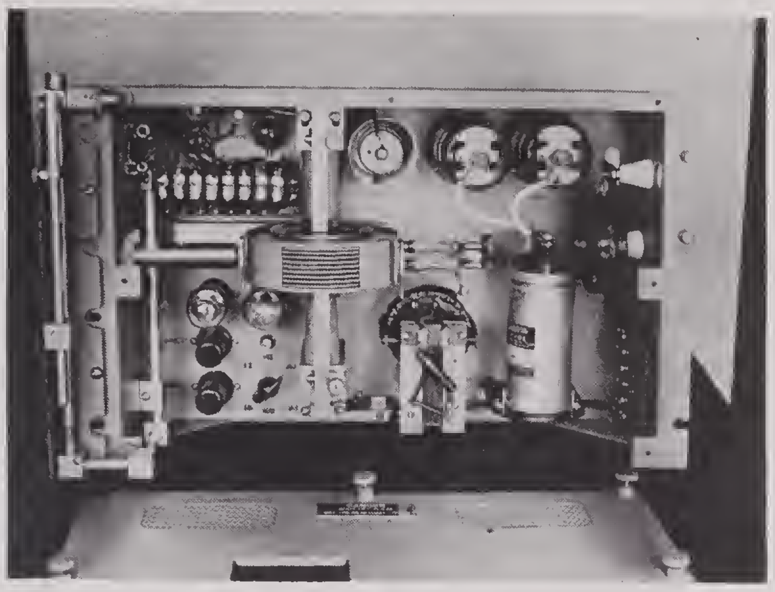

Figure 15. Interior view of pulse transmitter.

of which was determined by the dimensions of the tube. The magnetic field was supplied by permanent magnets of Alnico V and was adjusted by a variable shunt across each of the two magnets.

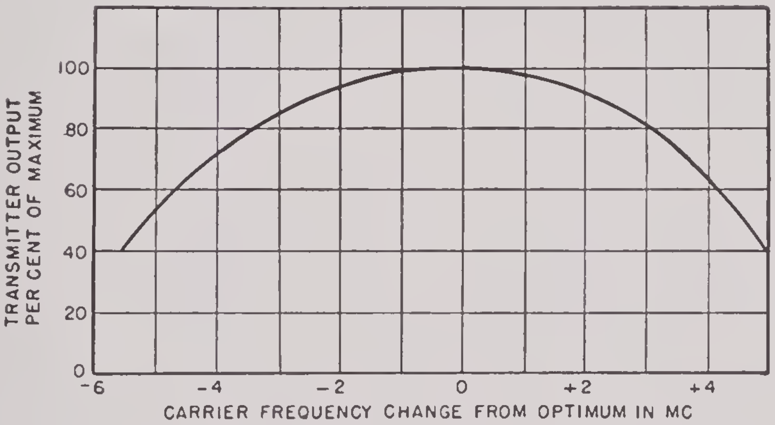

Figure 16. Effect on output of changing frequency by adjusting magnetron magnetic field and matching circuit.

The anode of the magnetron was at ground potential, excitation being provided by negative pulses applied to the cathode. These pulses were supplied by a pair of 813 beam-power tubes which were partially driven by regeneration from their own output and partly from a pulse oscillator using an 807 tube.

Table 2. Magnetron data.

Average power output	35 watts
Magnetic field	about 900 gausses
Average d-c current	26 ma
Peak pulse voltage	9,700 volts
Filament voltage	6.3 volts
Filament current	1 amp

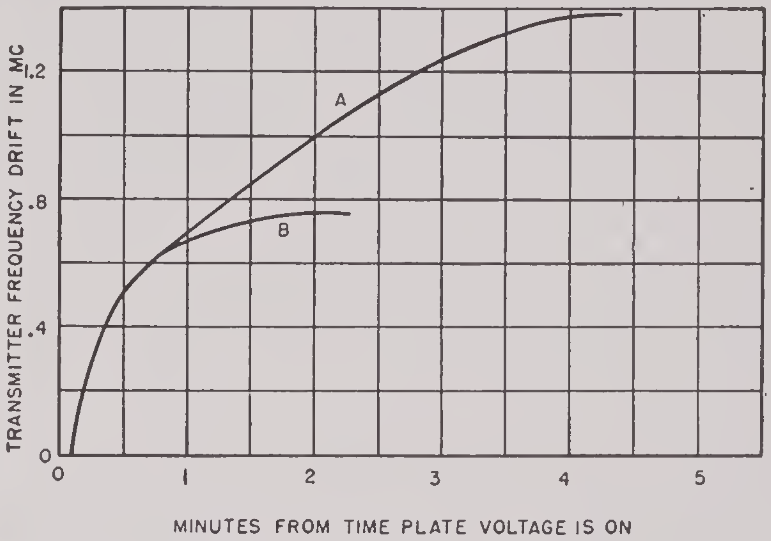

Figure 17. Frequency drift of transmitter. A, transmitter cold, filaments heated 1 minute; B, after 5 minutes stand-by period. Room temperature, 20 C.

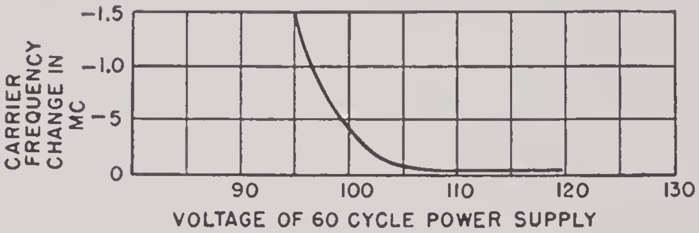

Figure 18. Relation between carrier frequency change and supply voltage changes.

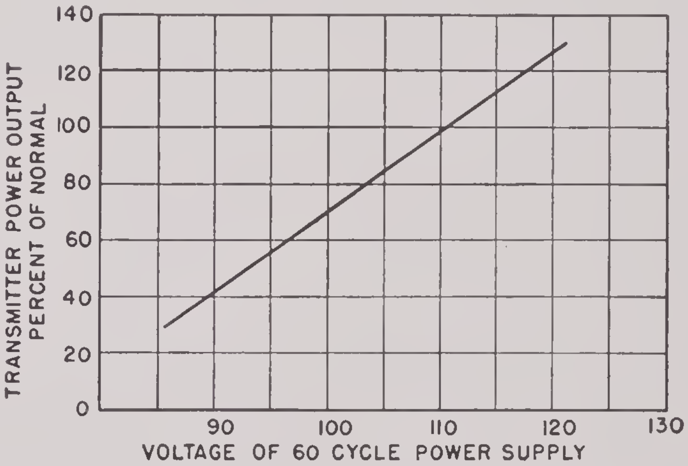

Figure 19. Effect on power output of changes in power supply voltage. At each voltage, magnetic field was adjusted for maximum output.

An artificial line with a delay from one end to the other of about 0.5 μsec, located in the cathode circuit of the 807, determined the pulse length. The frequency of the pulses was controlled by a 6SK7 frequency-modulator tube. Another 6SK7 and a 6H6 limited the maximum audio voltage applied to the control grid of the frequency-modulator tube.

The magnetron gave its greatest output at a particular frequency, and the output decreased if the load impedance and the magnetic field were changed to alter the frequency (Figure 16). The output frequency varied with changes in temperature (—13 cycles per million per degree C) and with line voltage (see Figures 17 and 18). The output power varied, of course, with changes in magnetron voltage as shown in Figure 19.

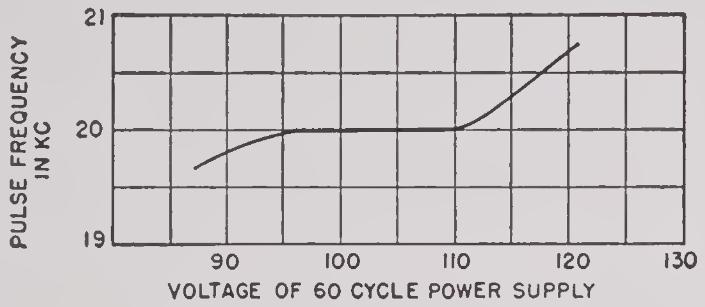

FIGURE 20. Effect on pulse frequency of changes in line voltage.

5.10.3 Transmitter Operation

Operation of the pulse transmitter (Figure 14) is as follows:

An increase of plate current of the 807 (V_1) through winding 1-2 of transformer T_1 induces a positive voltage on the control grid. This further increases plate current and results in a rapid rise of plate current through the tube to a high value.

The current flows from the cathode into the artificial line which during the current pulse is equivalent to a 152-ohm resistance (the characteristic impedance of the line) connected from cathode to ground. The voltage pulse produced by the current flowing into the line travels along the line to the open end, that is, to the end to which the anode of the 6SK7 is connected. This end is in effect open-circuited because of the high impedance of the 6SK7 compared to the line impedance. The voltage pulse is reflected without phase reversal and travels back along the line to the cathode. When the return pulse front, now at twice the original voltage, reaches the cathode, the cathode voltage rises and decreases the plate current. This causes a decrease in grid potential, which rapidly decreases the plate current to cutoff. The line will then be charged to a potential equal to about twice the cathode-to-ground potential at the beginning of the pulse. This charge will be reduced at a uniform rate by the plate current of the frequency-modulator tube V_2. When the cathode potential of the 807 reaches about 50 volts, the tube will again conduct current.

The rate at which the charge on the line is reduced and the value of the initial charge on the line determines the pulse frequency. The plate current of V_2 determines the rate of discharge of the line. The plate current is controlled by the screen potential and the cathode-to-grid potential. At the end of the pulse, the voltage charge on the line is proportional to the 500-volt supply. Thus if the supply rises in voltage it will be necessary to increase the plate current of V_2 to maintain constant pulse frequency. The plate current of V_2 is increased by increasing the screen-grid potential obtained from the 500-volt supply through R_4 and R_5.

Since it was necessary to by-pass the screen grid of V_2 for pulse frequency, C_2 was shunted across R_4 and R_5 so that the phase and amplitude of ripple frequency voltages would be maintained at their proper value at the screen grid. In other words, the time constant on R_4, R_5, C_2 was made equal to the time constant of by-pass capacitor C_3 and the screen grid resistance to ground.

Pulse frequency is modulated by voice frequencies by varying the grid voltage of V_2. A peak voltage of 2.2 volts on the control grid modulates the pulse frequency ±3 kc. The average pulse frequency is adjusted by R_4 to 20 kc.

AUDIO LIMITER

In order that modulation voltages applied to the control grid of V_2 should not exceed 2.2 volts, an audio voltage limiter is placed in series with the audio input. This limiter consists of V_3 whose amplification is controlled by the bias applied to its control grid by the left half of the 6H6 diode. The rate of decrease of this bias is governed by R_1C_1. This bias will drop to half value in about 1 second. The potential appearing across terminals 8 and 6 of the secondary of the modulation transformer is reduced by the divider R_2R_6. The volume control is set so that with maximum input from the microphone the peak voltage on the control grid of V_2 is 2.2 volts.

When the microphone switch is closed, the surge of current through the primary of the microphone transformer induces a high voltage in the secondary. The control bias resulting would keep the limiter cut off for several seconds and to prevent this from happening, the biased diode (right half of the 6H6) limits the maximum voltage which can be applied to the grid of V_3.

PULSE CONTROL

The pulse output of V_1 is amplified in V_5 (two 813 tubes in parallel) through T_2. Load current of the amplifier tubes flows through R_3 which is in series with the pulse autotransformer T_3 and across which is one winding of T_2. The voltage drop across R_3 is stepped up and reversed in polarity by T_2. The plus pulse voltage from oscillator V_1 starts the amplifier tubes conducting. The increase in plate current lowers the plate voltage which is again lowered by the increase of plate current due to the voltage across terminals 1 and 2 of T_2. The plate voltage thus drops to a few hundred volts. At the end of the pulse period the output of the oscillator drops to zero, reduces the plus voltage on the control grids of the amplifier tubes and their plate current increases. The regenerative action in T_2 causes the amplifier current to decrease rapidly to zero.

Sufficient fixed bias is supplied to the amplifier tubes to keep emission at zero or nearly zero in absence of pulse voltage excitation. During the operation the bias is increased by emission current so that the tubes are definitely cut off between pulses. This shortens up the output pulse and insures against passing any transients which may follow the pulse. Dimensions and windings of the pulse transformers are given in the final report[2] on Project C-42.

Output Circuit

Output power is indicated by a bolometer capacitively coupled to the output transmission lines. The bridge circuit, shown in Figure 21, is used to indicate a change of resistance of the bolometer resulting from heating by the r-f currents. The bolometer itself is a 5-ma vacuum fuse, one end of which is connected to ground through a section of coaxial line 1 in. long. This end is exposed to the r-f voltages. By acting as an inductance, the short length of line partially tunes out the capacitance to ground of the fuse cap. The other end of the fuse is by-passed to ground and is connected to the bridge through a flexible coaxial cable.

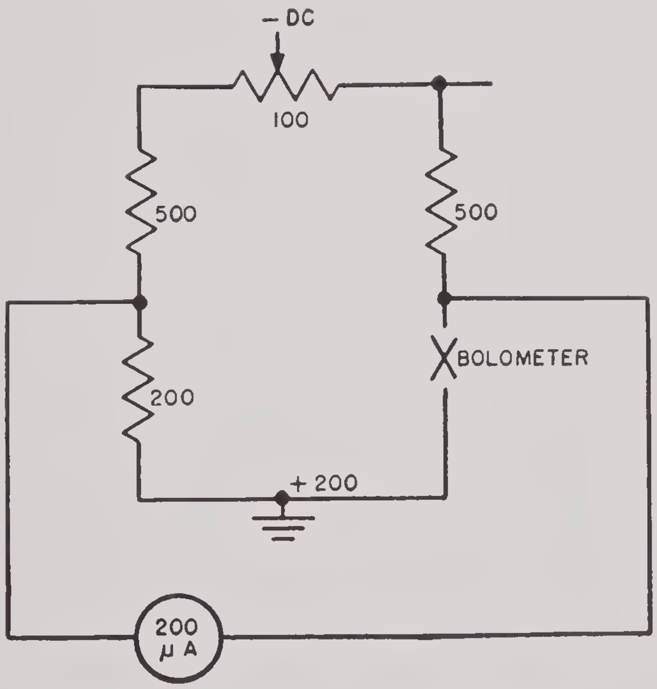

Figure 21. Circuit of bolometer bridge.

With no r-f voltage on the fuse the bridge is balanced and no current flows through the indicating meter. Radio-frequency voltages applied to the fuse change its resistance, upsetting the bridge balance. The resultant meter reading is proportional to the square of the r-f voltage. Polarizing voltage for the bridge is obtained from a 100-volt rectifier. The bolometer may be used to determine standing waves on the output transmission line or on the wavemeter to determine the output frequency.

Wavemeter

The wavemeter was a ¾-wavelength section of coaxial line. The length of the inner conductor could be varied 1 in. by a micrometer head at one end of the wavemeter, giving a frequency range of from 1,300 to 1,500 mc. At 1,400 mc a micrometer change of 1 division (0.001 in.) corresponds to a change of 0.220 mc of the resonant frequency.

5.11 RECEIVER

The receiver employed (Figures 22, 23, 24) covered a frequency range of 1,350 to 1,545 mc and consisted of an r-f amplifier, converter, and oscillator using GL-446 tubes followed by a 30-mc mid-band i-f amplifier having a band width of approximately 3 mc. The i-f amplifier was followed by a biased detector having a band-pass transformer in its plate circuit which converted received pulses into sine waves. The sine waves then passed through the limiter, discriminator, low-pass filter, and audio amplifier diagrammed in Figure 25.

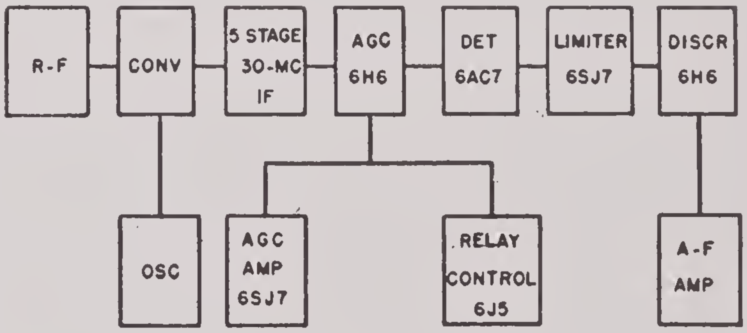

Figure 22. Block diagram of receiver.

The receiver had an excess noise ratio [ENR][d] of from 16 to 17 db ($kT\Delta f$ base) while the converter alone had an ENR of 21 db.

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[d] The excess noise ratio is a measure of the excess of the measured noise power output over the ideal noise power from the thermal agitation of the signal source.

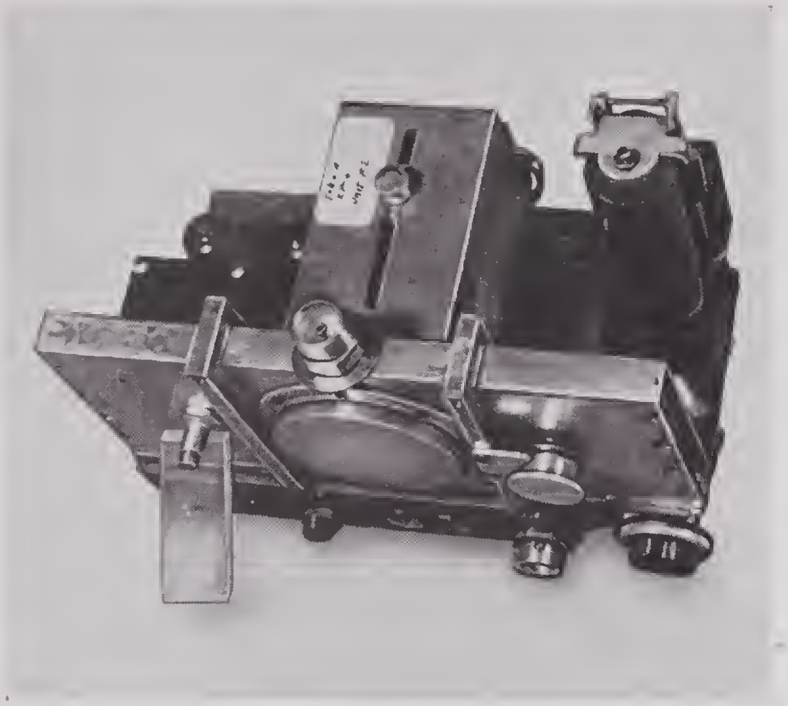


Figure 23. Radio frequency unit of directional receiver.

Oscillator

The oscillator frequency was lowered 340 kc from 1,480 mc (0.023 per cent) by a 10 per cent increase in line voltage and was lowered 1,280 kc (0.086 per cent) by a 10 per cent decrease in line voltage. Oscillator frequency decreased 780 kc from 1,470 mc for a 20 C increase in temperature, which is a temperature coefficient of —26 ppm/°C.

Intermediate-Frequency Amplifier

The converter was followed by a 5-stage 30-mc i-f amplifier, the first stages of which had automatic gain control [a-g-c] and the last two had fixed gain. If all 5 stages had a-g-c, there would not be enough voltage to operate the detector at high signal levels. The equivalent band width varied with received signal strengths due to the change in gain with a-g-c. The ENR of the i-f amplifier was 13.5 db.

Detector-Discriminator

The i-f amplifier was followed by a biased detector having in its plate circuit a tuned transformer with a pass band of 17 to 23 kc. This converted the received pulses into sine waves. The transformer fed a limiter tube which in turn fed a discriminator followed by a 4-kc low-pass filter to keep the 20-kc subcarrier out of the audio amplifier as well as to limit the audio band width to the speech range.

Automatic Gain Control

Automatic-gain-control voltage was obtained from a 6H6 connected across the output of the last i-f amplifier stage and was amplified by a 6SJ7 (Figures 26 and 27). The a-g-c voltage was also used to control the grid bias of a 6J5 which operated a carrier-alarm relay. The relay served two functions. When no signal was being received, it short-circuited one-half of the audio output transformer and prevented the noise level in the receiver output from rising too high. During a signal period, the short circuit was removed and a d-c voltage was supplied to the 600-ohm line to operate the alarm relay in a distant callbox.

Antenna

The antenna consisted of a vertical dipole and two high-impedance chokes to keep energy from traveling down the outside of the transmission line. The antenna was designed to match the 75-ohm line.

Callbox

The receiver had a relay operated by a-g-c voltage. In the absence of a signal this relay shorted half of the audio-output transformer and acted as a partial squelch. When a signal was received, the squelch was removed and a voltage to ground was applied to both sides of the handset-earphone line. This voltage was utilized to operate a carrier-on relay in the callbox which in turn could operate any sort of alarm circuit.

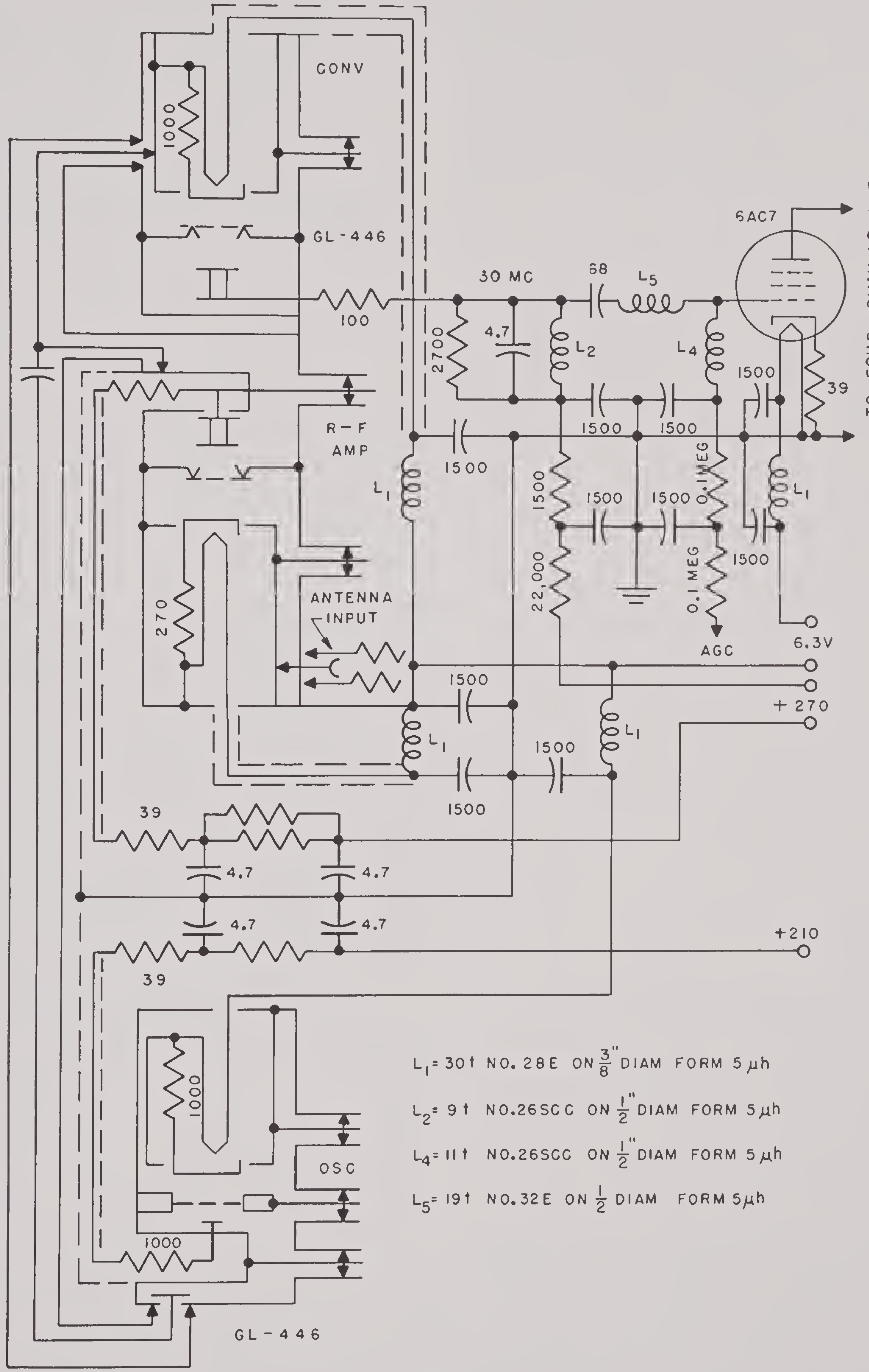
5.12 PERFORMANCE OF OVERALL SYSTEM

In addition to testing two of the units as a system, distortion measurements were made using both single tone and double tones. The effect of c-w or 1,000-cycle pulse interference on the receiver was investigated. Data on these tests are in the final report.[2]

Calculations were made on the probable range if one end of the circuit were in an airplane. As the distance is increased there is an area of continuous reception followed by an area where variable reception is obtained because of cancellation effects between the direct and reflected waves. As the distance is increased further a point is reached which is the limit of reception unless the antenna gain is increased. These signals could be received by an airplane at a distance of 124 miles if it were flying at an altitude of 1,000 ft.

5.13 DIRECTIONAL MICROWAVE TELEPHONE

A second part of Project C-42 was to develop a simple portable directional microwave communications



L_1 = 30 t NO. 28E ON $\frac{3}{8}$" DIAM FORM 5 μh

L_2 = 9 t NO. 26SCC ON $\frac{1}{2}$" DIAM FORM 5 μh

L_4 = 11 t NO. 26SCC ON $\frac{1}{2}$" DIAM FORM 5 μh

L_5 = 19 t NO. 32E ON $\frac{1}{2}$ DIAM FORM 5 μh

FIGURE 24. Input circuits to receiver.

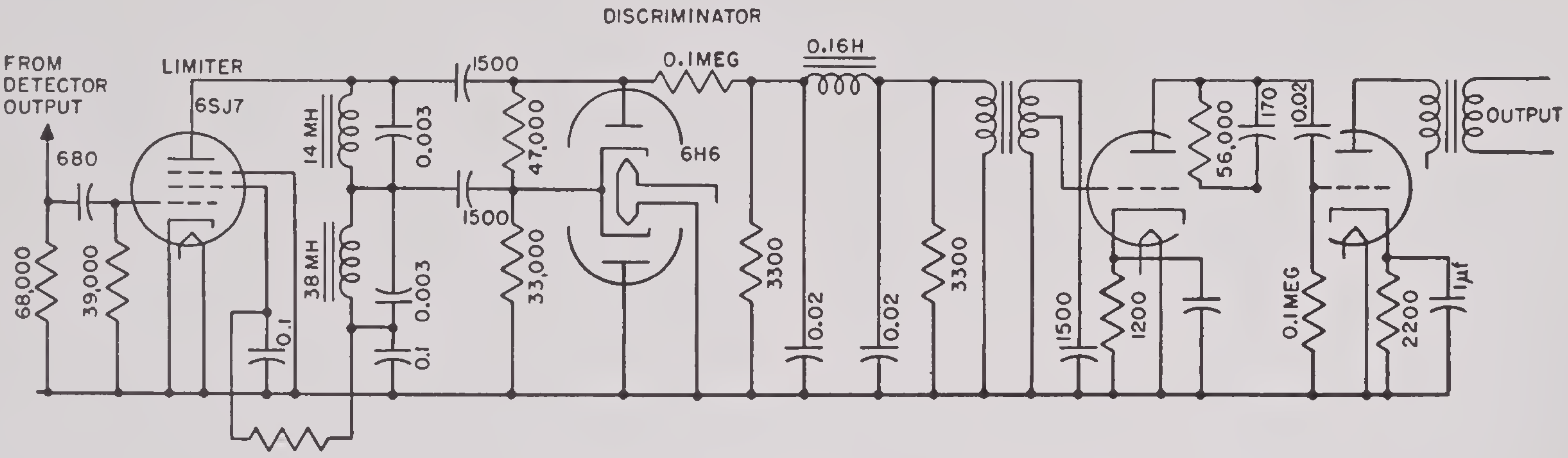


Figure 25. Circuit of limiter, discriminator, and output tubes.

system to replace visual signaling. The requirements were as follows: It should be light enough for two men to carry. It should have a minimum battery life of 4 hours, a maximum support height of 5 ft, a 20-degree beam, and provision for six channels. It should provide reliable communication with a clear line of sight over a 10-mile distance, simplicity of operation, security from interception, and suitability for operation in a net.

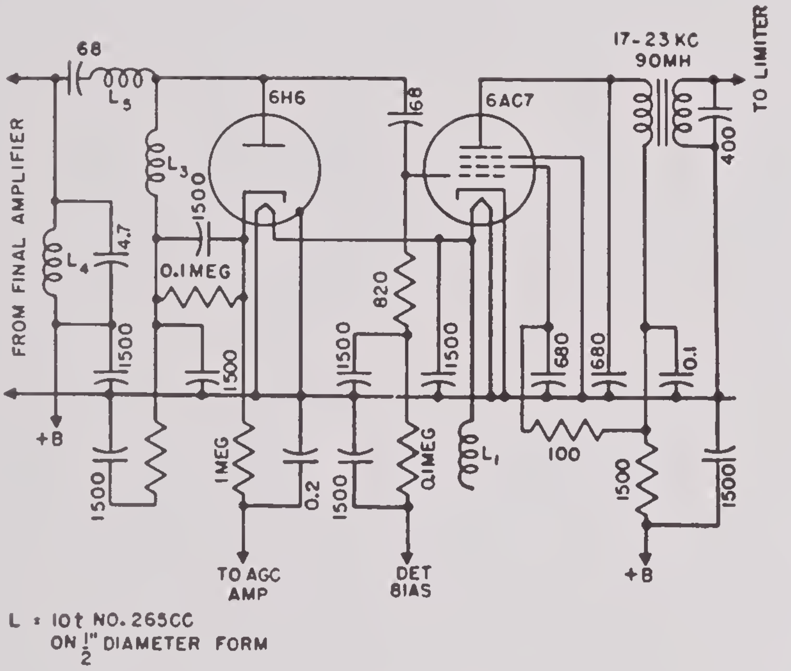


Figure 26. Circuit of a-g-c (6H6) and detector.

So far as these requirements are concerned these results were attained: Equipment was built into two packs, one weighing 24¾ lb and the other 31 lb. At 80 F about 6 to 8 hours of battery life could be expected which decreased to 4 hours at 20 F. The unit was mounted on standard Signal Corps tripods. (Figure 28 is a photograph of the system with antenna and reflector.) The beam was 6 db down at 12° total width and 20 db down at 26° total width. Consistent reliable communication was attained over distances of 30 miles from suitable elevations, and from favorable hills communication was carried on at 38 miles. The system had only one tuning knob and employed push-to-talk operation. Interception was difficult due to the narrow beam and to rapid attenuation beyond line of sight. The units could be adjusted by skilled personnel to any of 6 channels 4 mc apart. In the field, however, the transmitter tuning was fixed and the receiver could be tuned about ±2 mc from the transmitter frequency. Owing to the high directivity, it is doubtful that net operation would be possible. Units could operate on common frequency if separated about 50 ft and if the path in front of each unit were clear of reflecting objects.

5.13.1

Antenna System

The antenna was a λ/2 dipole and parasitic radiator housed in a plastic weatherproof box with a paraboloid reflector about 30 in. in diameter spun from sheet

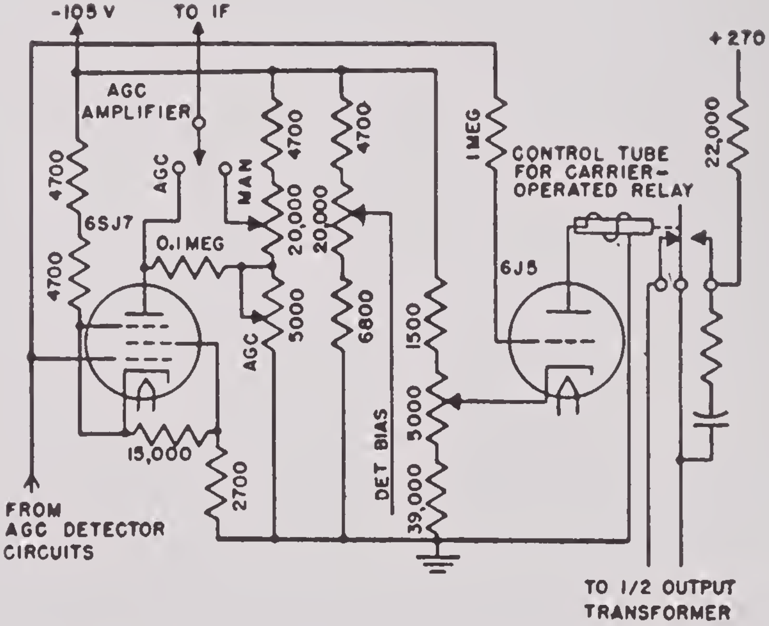


Figure 27. Circuit of a-g-c amplifier and callbox control tube.

aluminum. The dipole was fed by a concentric line with balanced feed obtained by a $\lambda/4$ sleeve at the end next to the dipole. A short piece of flexible concentric line connected the antenna to the oscillator.

5.13.2

Power Supply

The primary power source was three 2-volt, 25-amp-hr, lead-cell storage batteries (Signal Corps type BA-54-d, Willard-type radio 27-2). The cells weighed 4⅓ lb each. A standard Mallory VP-540 vibrator converted this energy to 250 volts at a maximum transceiver load of about 25 ma. The battery load was approximately 3 amp, receiving or transmitting.

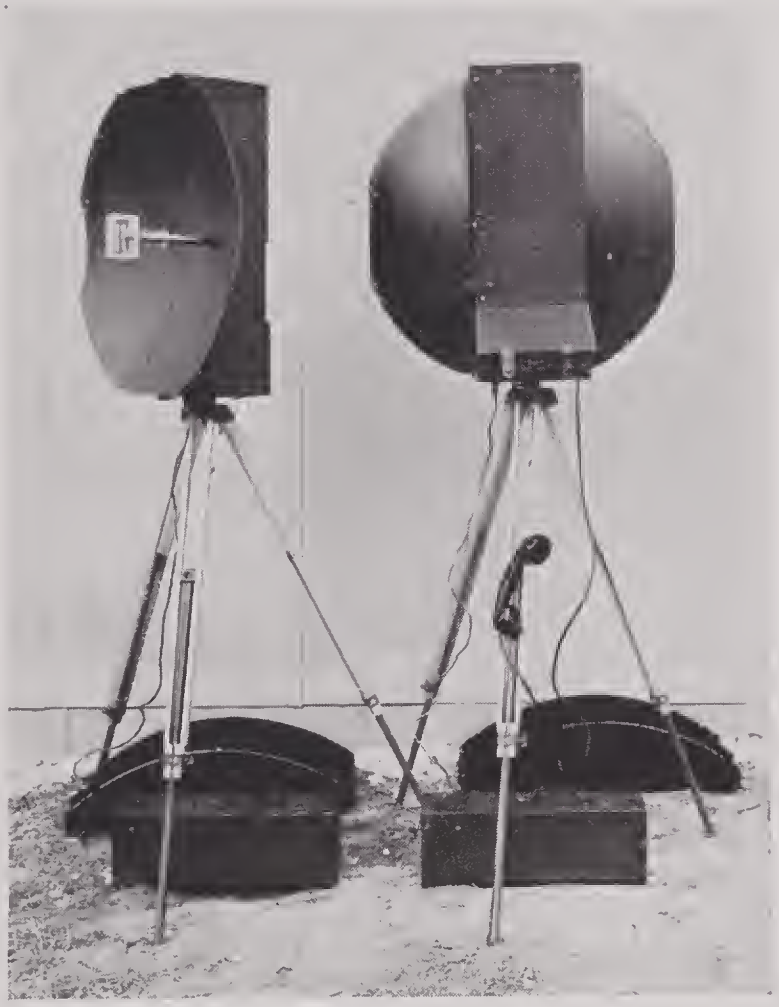


Figure 28. View of directional telephone system with antenna and reflector.

5.13.3

Radio-Frequency Unit

A cavity-type oscillator using a GL-446 lighthouse tube furnished output to the antenna. For transmitting the oscillator was cathode-modulated by a two-stage audio amplifier and for receiving the GL-446 was used as a superregenerative detector with separate quench oscillator and two-stage audio amplifier. The receiver had an ENR only 8 db worse than existing superheterodyne receivers using tube converters and no more than 16 db worse than the best receiver known at the time which used crystal converters. The ENR varied with signal strength because the a-g-c held the audio-frequency level down. Thus at 10-μv input the ENR was 32 db and with 500-μv input the ENR was 41 db.

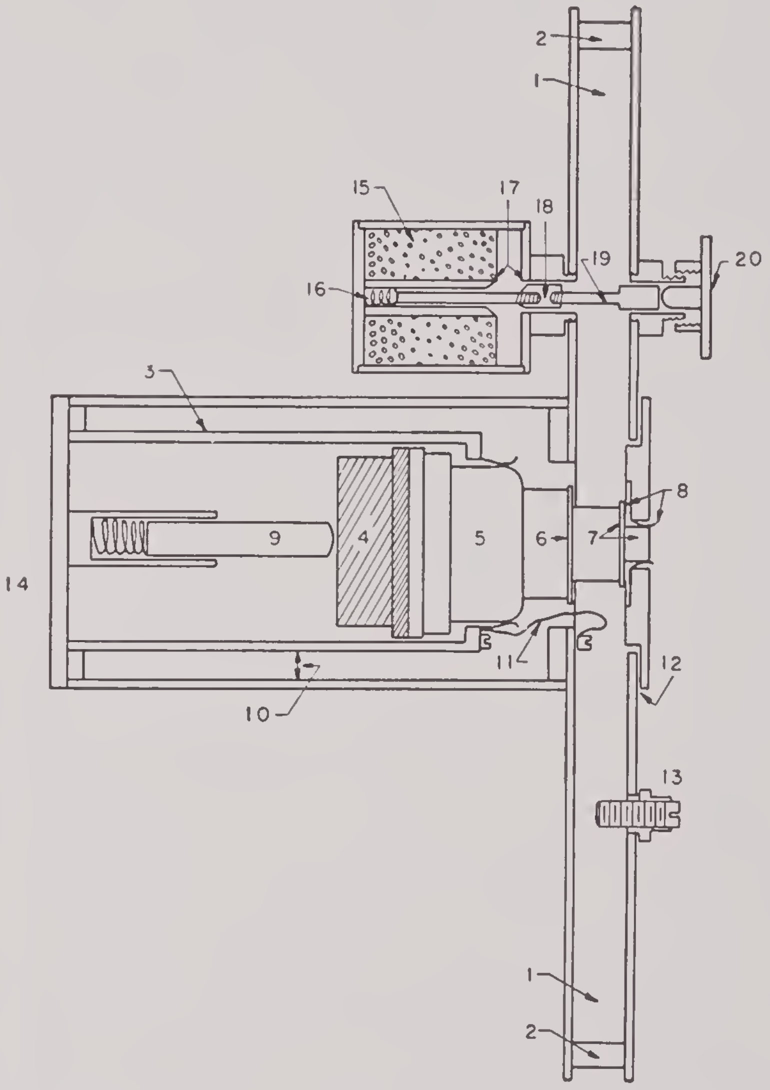


Figure 29. Cavity oscillator for directional system. 1, hollow cavity resonator operating at λ by $\lambda/2$ mode; 2, adjustable end plugs, set and clamped or soldered in in place; 3, inner conductor of short coaxial resonator carrying at one end spring fingers which contact tube shell, 5, of the GL-446 which is internally by-passed to cathode; 6, grid ring of GL-446 which contacts pair of 90-degree sector split contacts connected to left wall of cavity; 7, tube-plate ring which, through contact springs, 8, and mica by-pass ring, 12, connects to opposite wall. Thus coaxial, 3, in combination with slider, 10, tunes cathode-to-grid circuit, whereas cavity between grid and plate tunes that circuit. 9, spring-loaded plunger which holds tube against its spring contacts; 14, removable cover for changing tube; 11, small wire loop coupling plate grid to cathode grid circuit for feedback; 13, screw for setting channel frequency; 15 through 20, solenoid system for tuning receiver and compensating transmitter frequency to make it same as receiver frequency.

5.13.4 Transmitter

A minimum transmitter power of 100 mw was estimated to be needed with 20-db antennas at each end. Measurements on this unit as a transmitter showed 400 mw output at 5 watts input. Amplitude plate modulation up to 50 per cent was obtained but with concommittant frequency modulation of about 8 kc per volt. It was thought that CFVD modulation could be used to reduce the frequency modulation. A modulator made up of an 884 sawtooth generator operating at 20 kc, a 6SN7 pulse shaper, and a 6V6 modulator reduced the frequency modulation of the carrier, but actual listening tests showed no better results than with straight amplitude modulation. Since the CFVD system added several tubes it was abandoned. Further amplitude-modulation tests showed that at the same modulation capability could be secured with smaller modulator requirements by using cathode modulation. A 9002 triode was finally employed as a modulator. A quench output of 100 volts at 100 kc was decided on.

Measurements indicated that the transmitted and received frequencies of the oscillator would differ by 0.5 to 1.5 mc. To overcome this common transceiver fault a compensating device was added to the cavity. This is illustrated in Figure 29.

5.14 FIELD TESTS

The longest path over which communication was held was along Long Island Sound over salt water. One terminal was on a 225-ft hill and the other on a 120-ft hill. Total distance was 38 miles and the visibility was calculated to be 36 miles based on $\frac{4}{3}$ earth's radius. This path showed some fading, as would be expected for a path exceeding the optical range. Other tests gave good communication at distances of 29 miles. Communication was also carried on over water with heights of about 4 ft and 15 ft, with the transceivers on a beach 5.75 miles apart.

Chapter 6

R-F GENERATOR FOR 2,000 TO 3,000 MC

A pulse-modulated telephone generator tuning from 1,980 to 3,120 mc, having a frequency stability better than ±0.05 per cent over an ambient temperature range of ±10 C, producing approximately 10 watts, using a velocity variation type of tube.

6.1 INTRODUCTION

At the time this project[a] was started there was no available equipment for producing voice-modulated power of several watts in the 2,000- to 3,000-mc range, with provision for frequency change over so wide a range and meeting the requirement of 0.1 per cent in frequency constancy at any selected operating frequency.

Under the project a model was developed the oscillator of which is tunable from 1,940 to 3,150 mc, although the frequency range of the entire unit is limited by the tunable range of the monitor cavity which is 1,980 to 3,120 mc. Modulation up to 100 per cent is effected by varying the duration of superaudible 16-kc pulses. The tube used as oscillator is of the velocity variation type, having a focused beam and using an external magnetic field. This tube was selected because it had reached the point where commercial application was feasible. Its construction permits the use of an external circuit so that wide tuning range is possible. Other tubes available at that time did not permit such a wide range or would not deliver so much power.

6.2 THE OSCILLATOR TUBE

The heart of the transmitter is the oscillator tube (1290-CT) with its associated cavity, two walls of which are adjustable by dial controls to permit tuning to the desired frequency. By splitting the cavity along its center plane the two halves can be pulled apart so that the tube can be removed and replaced easily.

The output coupling system consists essentially of a short piece of coaxial cable projecting into the cavity and provided with a small coupling loop on its inner end. The orientation and location of the loop within the cavity may be adjusted from outside to vary the coupling between cavity and external load.

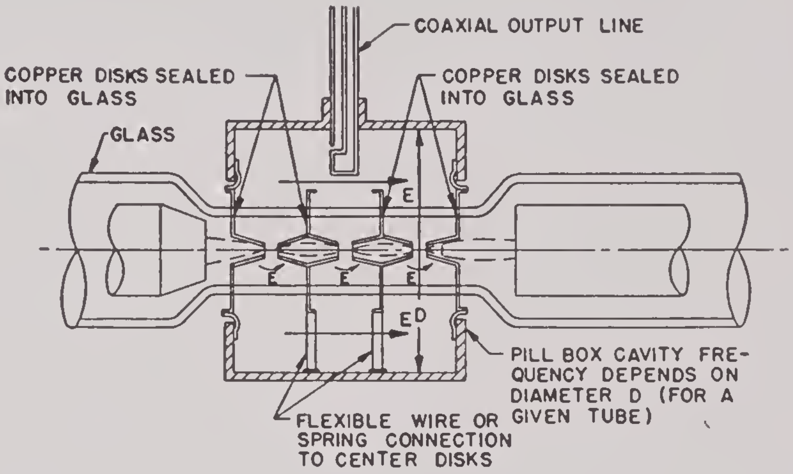

COAXIAL OUTPUT LINE
COPPER DISKS SEALED INTO GLASS
COPPER DISKS SEALED INTO GLASS
GLASS
E
E
E
E
E
D
PILL BOX CAVITY FREQUENCY DEPENDS ON DIAMETER D (FOR A GIVEN TUBE)
FLEXIBLE WIRE OR SPRING CONNECTION TO CENTER DISKS

Figure 1. Schematic drawing of "three-gap" velocity variation oscillator tube with cavity.

The cavity is made of rectangular bronze tubing 4x1¾ in. outside dimensions and 7 in. long. Two plungers are employed having the shape shown on the left of Figure 4. The tuning range of the cavity with a "three-gap" tube is shown in Figure 5. Some trouble was had with parasitic oscillations because the connections to the two center disks were brought out through holes in the cavity. The arrangement shown in Figure 6 cured this difficulty. The leads from the two disks form a transmission line terminated at one end by the disks and at the other by a small capacitance equal to about 1 $\mu\mu$f between the two leads. Resistors connected across this capacitance are isolated from the 10- to 15-cm oscillations by small choke coils. They stop any tendency to oscillate at about 40 cm but have no effect on the desired oscillations. The length of the transmission line is adjusted by building into the cavity with a small copper block so that the line length lies between $\lambda/2$ and $\lambda/4$ throughout the operating range.

6.3 MAGNETIC FIELD

The field required for focusing the three-gap tube must be nearly uniform throughout the space occupied by the beam and must be variable between about 400 and 600 gausses. This field is supplied by two field coils with a soft iron shell. The requirements on ripple suppression were not so severe as those imposed upon

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[a] Project C-7; Contract NDCrc-177, Western Electric Co., Inc. The generator described herein was supplied to RCA Communications, Inc., for field tests covered in Chapter 5 under Project C-24.

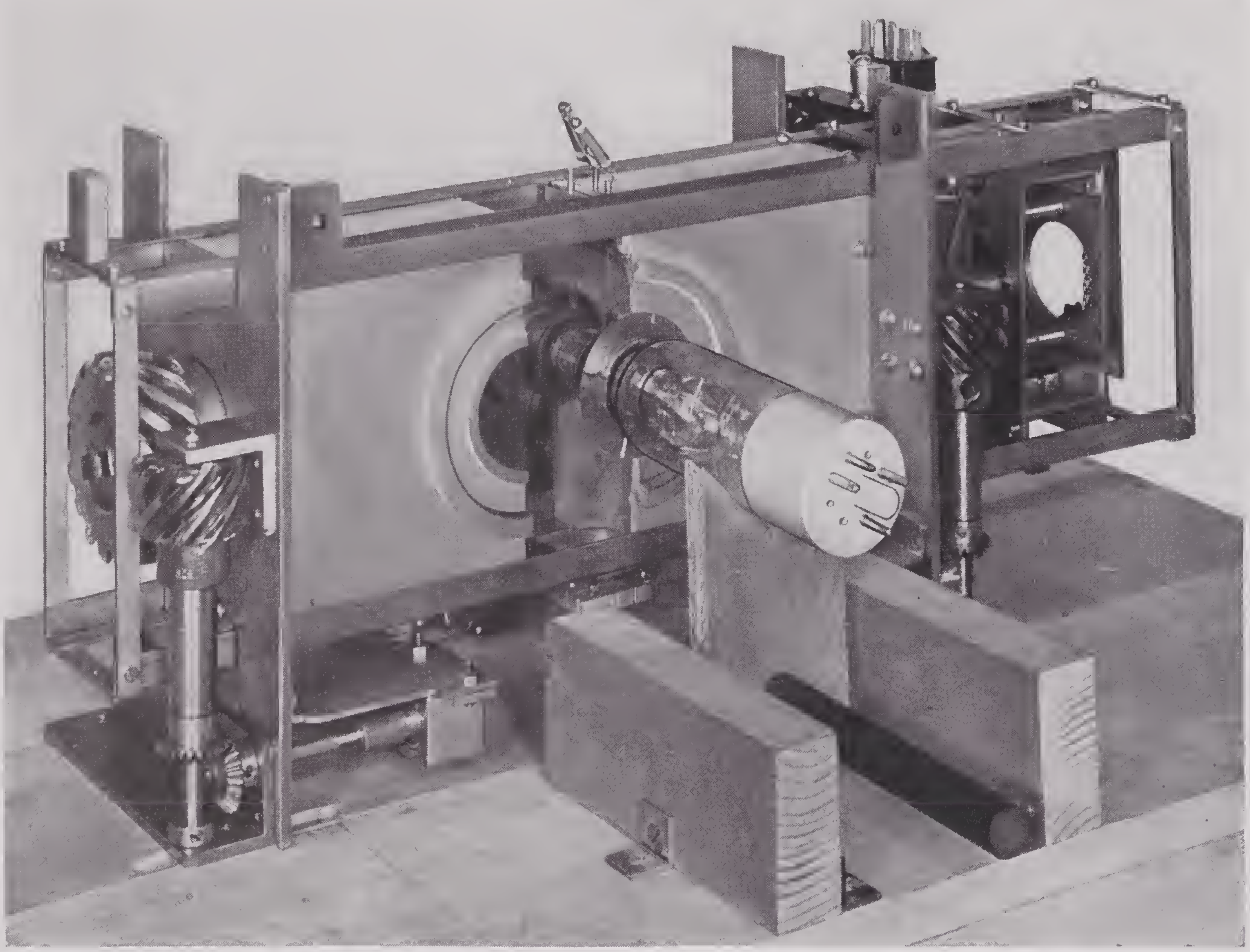

FIGURE 2A. Photograph of oscillator tube showing demountable cavity.

the supply to the tube disks, so that no electronic voltage regulator was required on the power supply for the field.

6.4 AUTOMATIC TUNING SYSTEM

Early tests showed that considerable frequency variation would be experienced during the warm-up period of the oscillator cavity and tube. Variations of about 5 mc at 10 cm and 2 mc at 15 cm were encountered, indicating that some means of automatic tuning would be necessary to hold the transmitter within the specified frequency.

The final arrangement consists of a ½-in. rod made of thin-walled brass tubing arranged to be driven into and out of the cavity by means of a reversible motor mounted on a carriage at one end of the cavity structure. The motion of the tuning plunger was set at 1 in. At 10 mc, however, this motion would have placed the plunger beyond the glass of the tube, so the carriage carrying the motor was arranged to be retracted as the cavity piston approached the tube. The actual means by which the frequency is controlled by this plunger will be evident from a description below.

6.5 MONITOR CAVITY

The standard of frequency about which the transmitter is designed to operate lies in the small so-called cavity monitor, consisting essentially of a small wavemeter tuned by means of a micrometer head (Figure 7). This wavemeter consists of a coaxial transmission line shorted at one end and open at the other the length of which is adjustable to $\lambda/4$ at the desired frequency by means of the micrometer. A small silicon crystal rectifies the r-f output of this cavity, the rec-

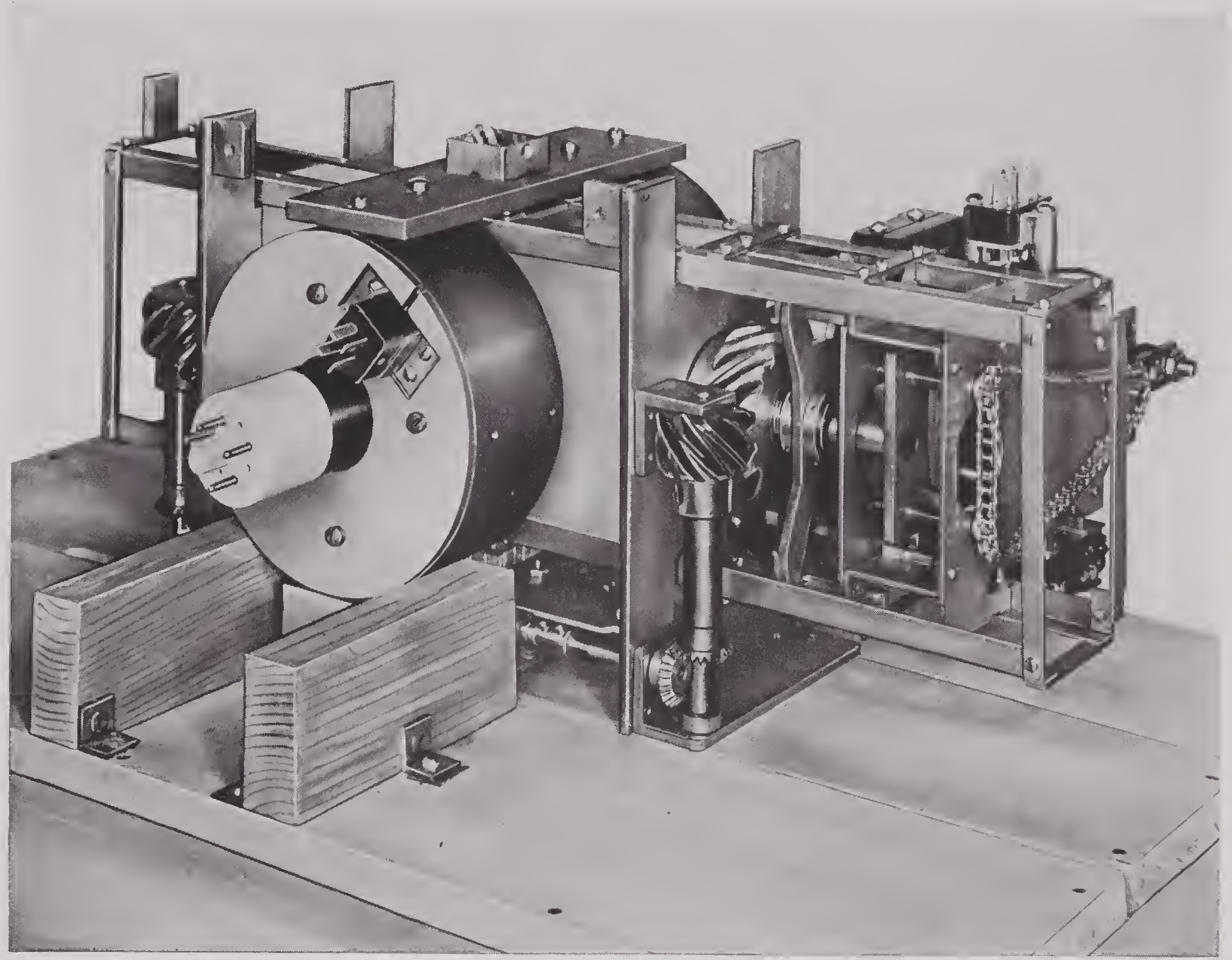

Figure 2B. Photograph of assembled oscillator tube.

tified crystal current being used for a side-tone circuit and for the automatic tuning system.

A rotating cam enters a slot in this cavity and varies the resonant frequency at a rate corresponding to about 80 cycles per second. Thus if the mean frequency of the transmitter is on one side of the monitor-resonance curve, the result will be an 80-cycle modulation of the output of the cavity crystal. If the transmitter is on the other side of resonance with the cavity, the 80-cycle modulation will be reversed in phase. If the transmitter is at the top of the resonance curve, modulation current will vanish. Thus it is only necessary to amplify the 80-cycle crystal output and to commutate it synchronously, whereupon the direct current will be zero at resonance and will have a positive or negative direction depending upon which side of the resonance curve the transmitter is operating. An electric filter and a polarized relay controlling the direction of motion of the automatic tuning motor mounted on the oscillator cavity structure, and therefore controlling the position of the tuning rod in the transmitter cavity, make up the remainder of the tuning system.

A two-stage amplifier made up of a 6SJ7 and a 6Y6G tube provides sufficient gain for the automatic tuning system even when only one-quarter of the normal current is present in the monitor cavity. This system holds the average frequency of the transmitter to within ±300 kc of the frequency of the monitor. The range over which the transmitting frequency can be varied by the tuning motor is about 7 mc at 15 cm and 25 mc at 10 cm. This is sufficient to cover the variations during the warm-up period, to allow for changes in tubes, and for variations in the output coupling loop and in the magnitude of the disk voltage.

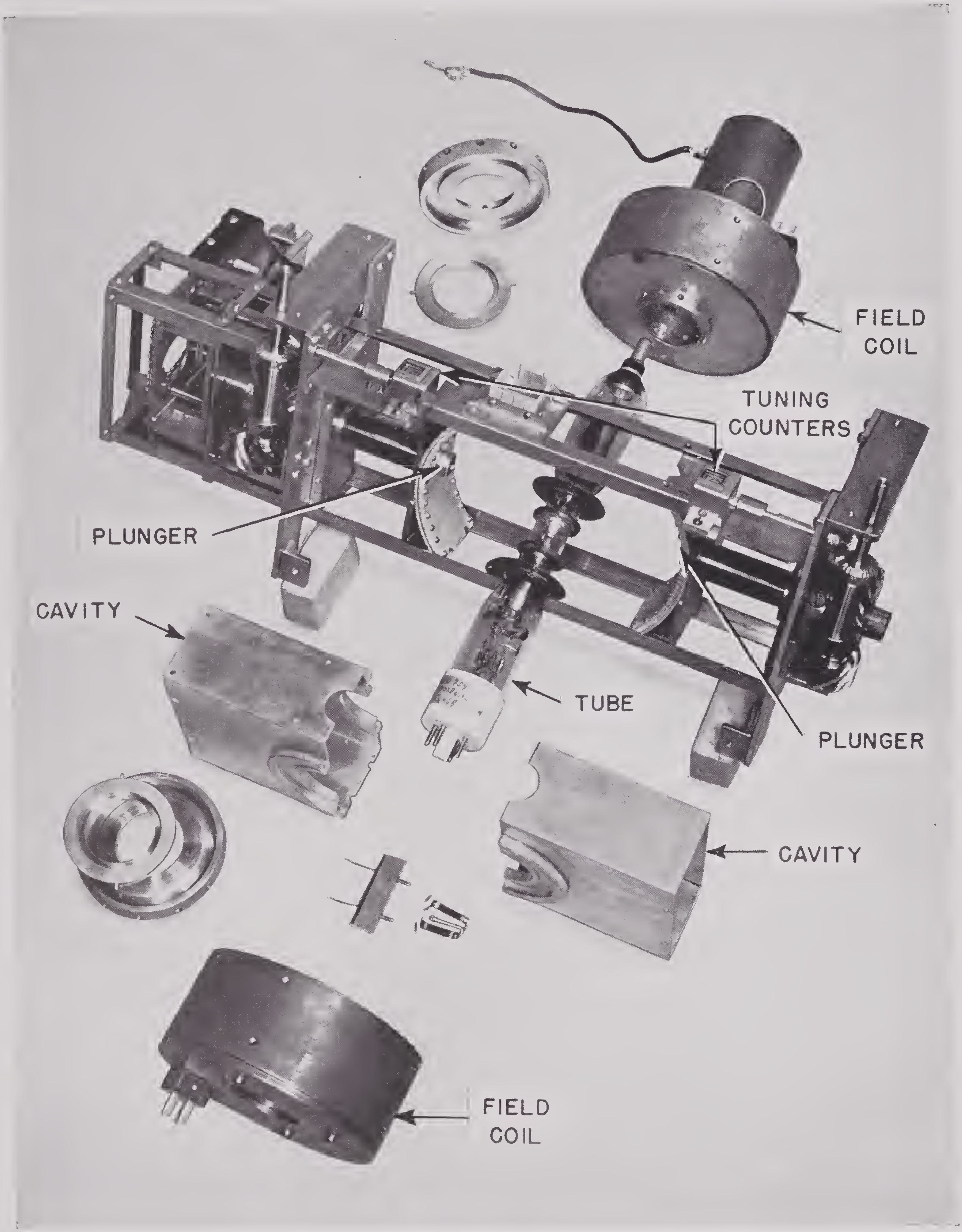

FIGURE 3. Knocked-down velocity variation tube employed in voice-modulated pulse transmitter.

CONFIDENTIAL

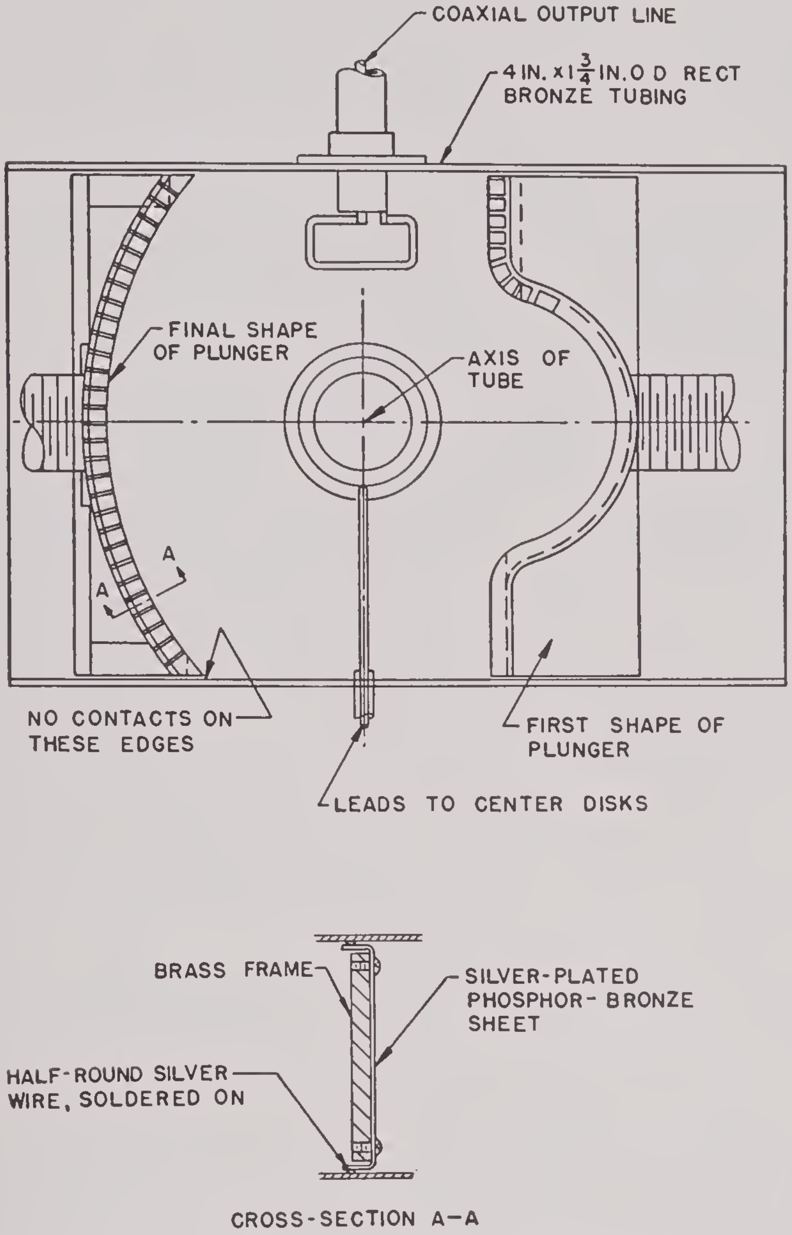
COAXIAL OUTPUT LINE
4 IN. x 1¾ IN. O D RECT BRONZE TUBING
FINAL SHAPE OF PLUNGER
AXIS OF TUBE
A
A
NO CONTACTS ON THESE EDGES
FIRST SHAPE OF PLUNGER
LEADS TO CENTER DISKS
BRASS FRAME
SILVER-PLATED PHOSPHOR-BRONZE SHEET
HALF-ROUND SILVER WIRE, SOLDERED ON
CROSS-SECTION A—A

Figure 4. Details of two-plunger cavity showing (on left) construction of plunger finally used.

6.6 CFVD MODULATION SYSTEM

Various methods of modulating the velocity variation tube were considered (described in the final report[1]) with the result that the following scheme was selected and put into the model transmitter.

This method consists, briefly, in modulating the length of square pulses of constant amplitude and frequency. One advantage is the fact that special forms of limiting can be used at the receiver that cannot be used with amplitude modulation. With a combination of modulation voltages on the accelerating anode and on the disks the best linearity and frequency modulation compensation can be obtained. In this way it is possible to reduce the frequency modulation from about ±1 mc to 100 kc, but the addition of more filter stages to the disk voltage supply further reduces ripple voltage to about 85 db below the d-c output, with the result that frequency modulation due to ripple voltages is of the order of ±10 to 50 kc. Use of a-c heater voltages brought this undesired frequency modulation up considerably and was abandoned in favor of battery supply. A circuit diagram of the continuous-frequency variable-duration [CFVD] pulse transmitter is given in Figure 9.

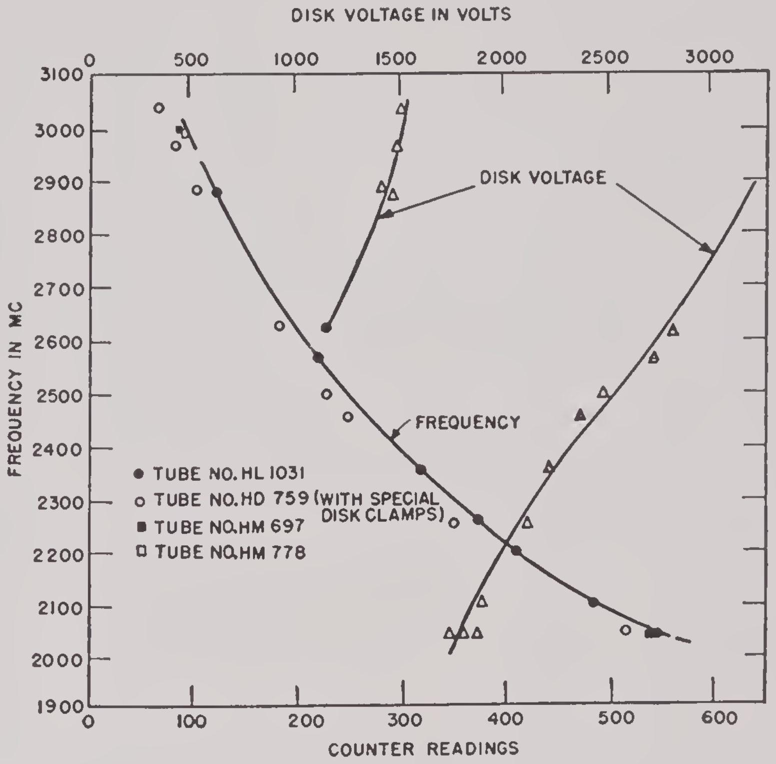

Figure 5. Approximate frequency calibration of oscillator.

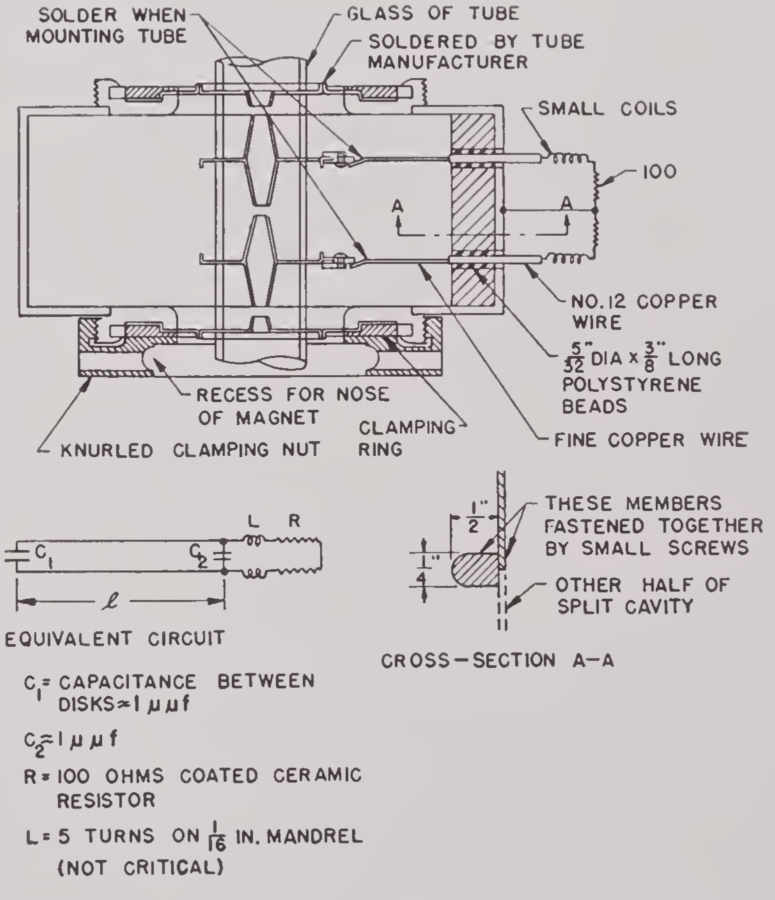
SOLDER WHEN MOUNTING TUBE
GLASS OF TUBE
SOLDERED BY TUBE MANUFACTURER
SMALL COILS
100
A
A
NO. 12 COPPER WIRE
5/32" DIA x 3/8" LONG POLYSTYRENE BEADS
FINE COPPER WIRE
RECESS FOR NOSE OF MAGNET
CLAMPING RING
KNURLED CLAMPING NUT
L R
C1
C2
ℓ
EQUIVALENT CIRCUIT
C1 = CAPACITANCE BETWEEN DISKS ≈ 1 μμf
C2 = 1 μμf
R = 100 OHMS COATED CERAMIC RESISTOR
L = 5 TURNS ON 1/16 IN. MANDREL (NOT CRITICAL)
1/2"
1/4"
THESE MEMBERS FASTENED TOGETHER BY SMALL SCREWS
OTHER HALF OF SPLIT CAVITY
CROSS—SECTION A—A

Figure 6. Details of cavity and equivalent circuit.

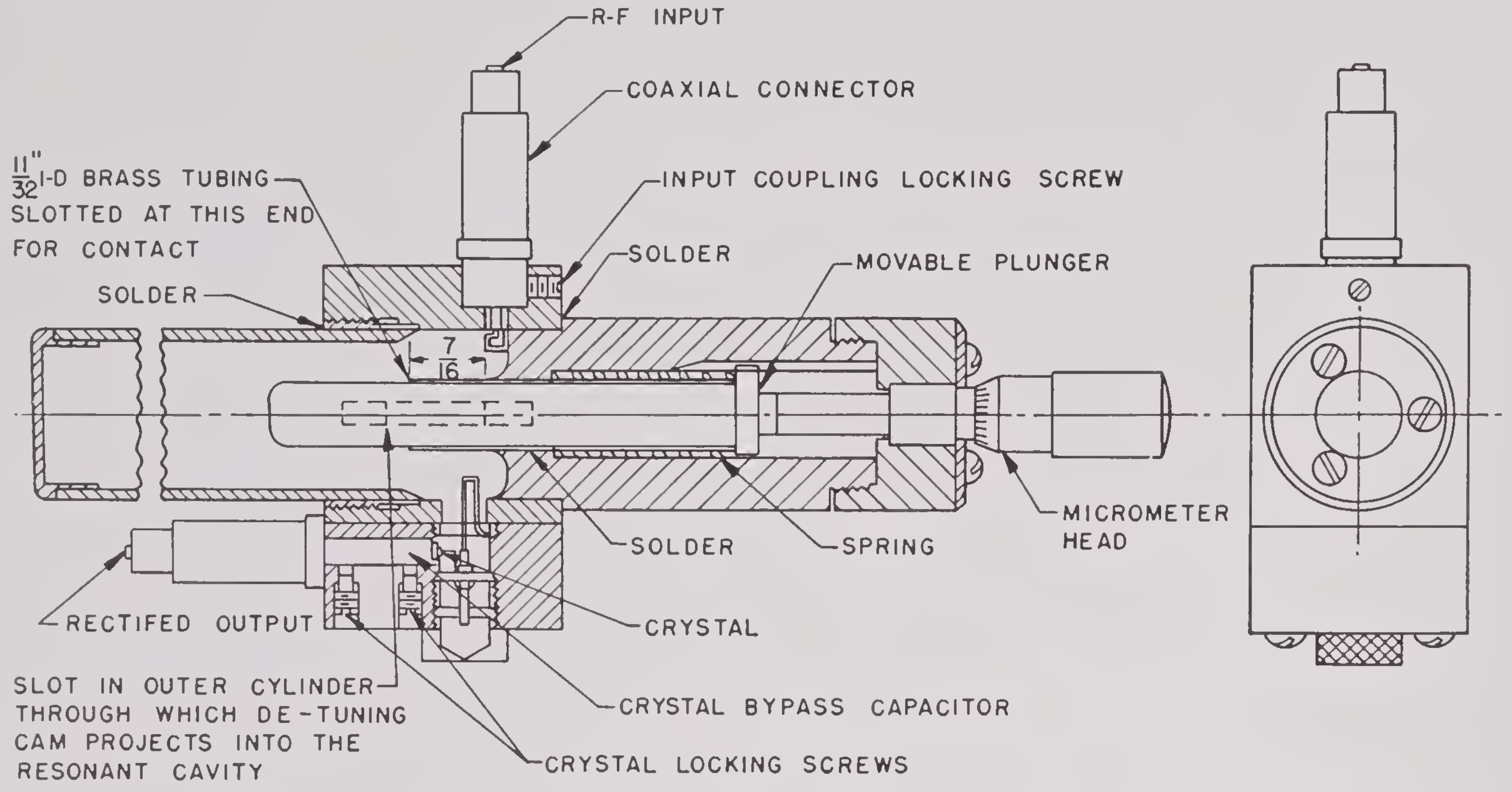
R-F INPUT
COAXIAL CONNECTOR
$\frac{11}{32}$" I-D BRASS TUBING SLOTTED AT THIS END FOR CONTACT
INPUT COUPLING LOCKING SCREW
SOLDER
MOVABLE PLUNGER
SOLDER
$\frac{7}{16}$
SOLDER
SPRING
MICROMETER HEAD
RECTIFED OUTPUT
CRYSTAL
SLOT IN OUTER CYLINDER THROUGH WHICH DE-TUNING CAM PROJECTS INTO THE RESONANT CAVITY
CRYSTAL BYPASS CAPACITOR
CRYSTAL LOCKING SCREWS

Figure 7. Mechanical details of 10- to 15-cm wavemeter used as monitor for obtaining automatic frequency control.

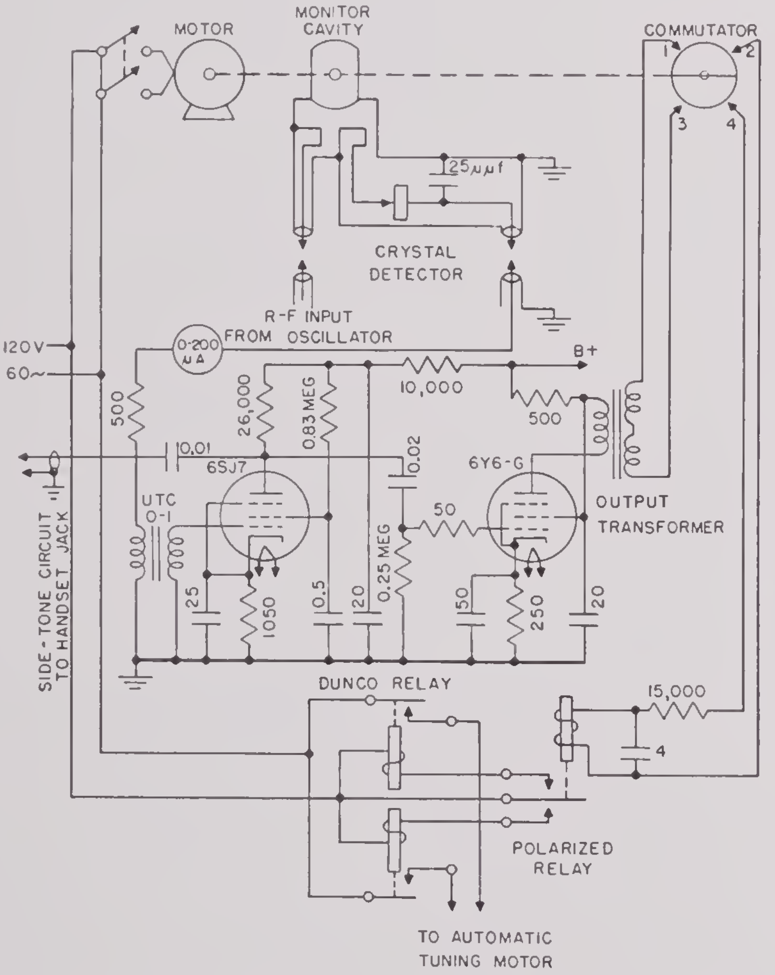

Figure 8. Circuit of frequency monitor and amplifier.

6.7 PULSE GENERATOR

The pulse generator consists of a multivibrator generating an approximation to a square wave at about 16 kc followed by a two-stage pulse shaper and a series-integrating circuit made up of capacitance and resistance which produces a voltage wave approximately triangular in shape. The distance along a horizontal line between two successive points on the triangular curve is almost exactly proportional to the height of the horizontal line above the axis. Thus it is only necessary to apply this wave to the input of a tube operating around cutoff and vary the bias on the tube to correspond with the modulation desired.

The triangular wave and the audio voltage may, ordinarily, be applied to the modulator tube without excessive interaction but in this case the small ratio between the frequencies of the triangular wave (16 kc) and the highest audio frequency make the problem somewhat more difficult. By means of a cathode follower connected directly to the cathode of the modulator the triangular wave voltages appear across a low impedance, with the result that the triangular wave is practically independent of the cathode current. Audio modulating voltages are applied directly to the grid as a bias variation.

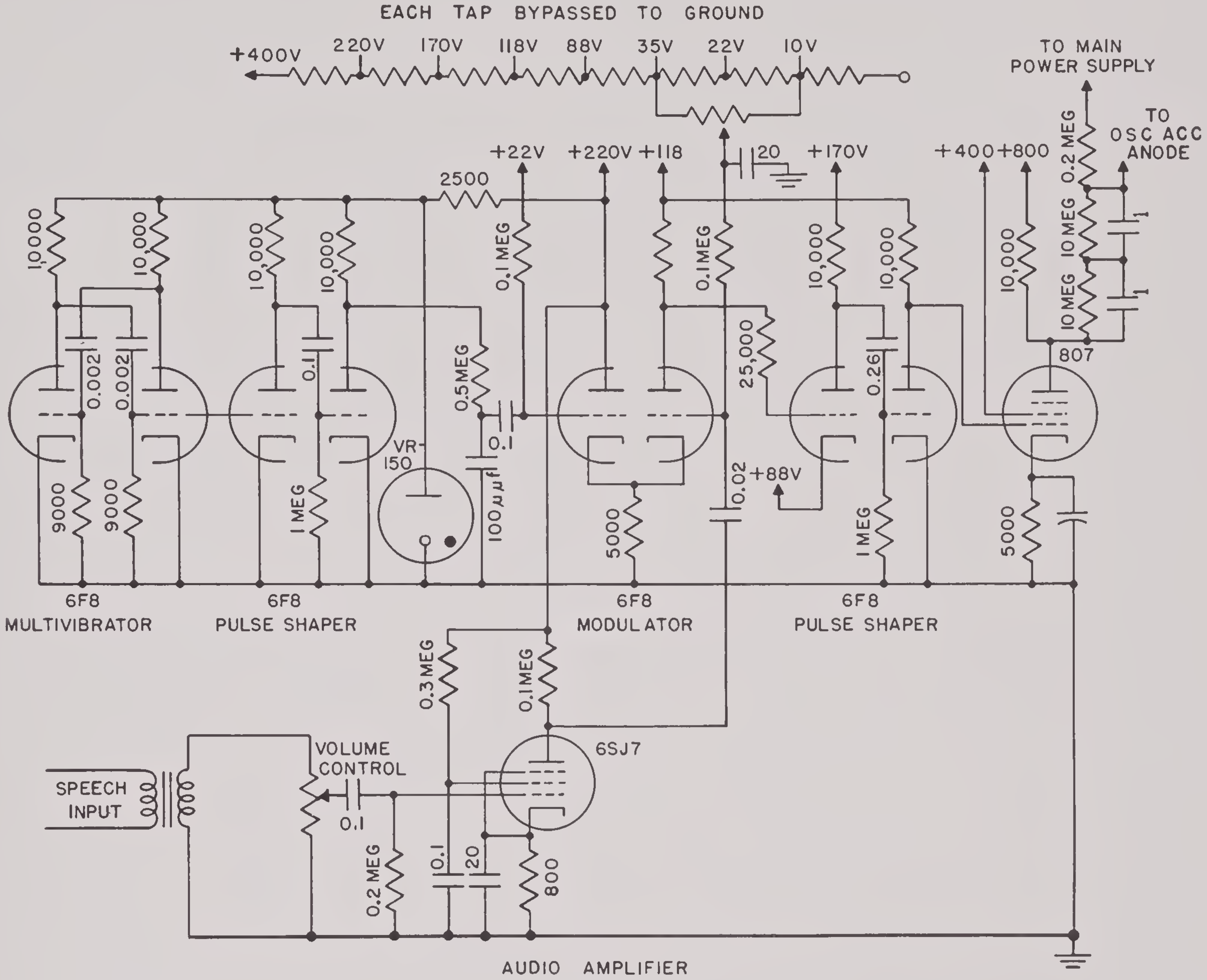

FIGURE 9. Circuit diagram of continuous-frequency variable-duration [CFVD] pulse transmitter.

Following the modulator is a two-stage pulse shaper which provides an output of 60 volts peak-to-peak, the shape of which is nearly rectangular. This voltage drives an 807 output tube.

During a cycle of modulation the duration of the pulse varies from very nearly zero to a complete cycle, thus varying the average power output of the tube. Measurements indicated that both second and third harmonic distortion in this system are down about 20 db at 100 per cent modulation.

This system of modulation was chosen after preliminary experiments indicated that the modulation characteristic of a velocity-variation type of oscillator was exceedingly nonlinear and that there was considerable spurious frequency modulation. Even in the case of the modulation employed (CFVD) a certain amount of frequency modulation exists because of the fact that the beam in the tube does not reach its final condition until about 5 μsec after it has been turned on. Therefore during the first 5 μsec of each pulse the frequency of the transmitter differs somewhat from the frequency during the remainder of the pulse. The difference at 10 cm may be as much as 2 mc but decreases at longer wavelengths becoming negligible at 15 cm. This frequency modulation does not occur at voice frequency rates as in some modulation systems.

6.8 POWER SUPPLY

The main power-supply system is capable of delivering any desired voltage between 1,500 and 3,700 volts and is electronically regulated to maintain low ripple voltages. The electronic regulator requires less space and weight than the chokes and capacitors which would be required to reduce the ripple an equivalent amount.

PART III

PRECIPITATION STATIC PROBLEMS

With the inauguration of long-range military flights incident to the problems of national defense and later of actual warfare, which had to be carried out in areas of prevailing bad or adverse weather conditions such as Alaska, Northwestern United States, etc., the radio interference phenomenon known as precipitation static became of primary concern and Division 13 undertook the prosecution of a general study. Precipitation static is caused by high-voltage electric discharges from planes due to the accumulation of charges picked up by flying through snow, rain, or dust. These accumulations raise the airplane to a potential which will break down the insulating ability of the surrounding atmosphere. When encountered, the interference caused usually entirely disrupts radio communications.

Several contracts were entered into by Division 13 on various phases of the project. It was soon found that military operation of aircraft at much higher altitudes and at much greater speeds resulted in greatly increased interference which could not be combated effectively by the means used by the commercial air transport operators. Further investigation with the hope of better solutions was required. Furthermore the hazard of lightning strikes to aircraft in flight constitutes a problem closely related to precipitation static and was made a subject for concurrent investigation.

Several of the studies were conducted entirely in the laboratory, while others employed both laboratory techniques and flight tests to examine the usefulness of the instruments developed and flight-tested. The progress of the project as a whole was delayed for some time due to the unavailability of aircraft having the required size and speed characteristics for carrying forward the flight-test program.

Chapter 7

PRECIPITATION STATIC REDUCTION

Investigation of the fields around point-discharge sources; effects of air movement and radiation on point discharge; shock excitation of radio receivers; noise reduction by nonlinear devices; charge dissipators, such as fine wires, the Bendix discharger, radioactive devices, and flame dischargers.

7.1 INTRODUCTION

Under this project[a] several aspects of the precipitation static problem were studied.[1] Among these studies are the mechanism of high-voltage electric discharges, the approximate equivalent circuit for studying voltage-induction effects arising from corona discharges, the statistical current-frequency spectrum of the positive streamer discharge and the determination of the corresponding current-burst shape, determination of pulse rates for negative-point discharges, radio interferences to be expected in complex networks, effect of receiver detectors on measured interference, shock-excitation effects and the relationship between effective band width and circuit decrement, and noise reduction secured by means of limiters, dampers, and other nonlinear circuit elements.

7.1.1 Apparatus Involved

To supply the high voltages needed to produce electric discharges at atmospheric pressure, rectifier-filter arrangements were provided producing 200,000 volts and discharge currents of 0.005 to 5,000 μa. Means were also provided for superimposing on the d-c potentials an oscillating voltage of several thousand volts and of any frequency between 100 and 4,000 cycles per second. Vacuum-tube voltmeters were employed for measuring interference.

7.1.2 Preliminary Investigations

Before studying actual means for reducing precipitation static noises, an extended study was made of various types of noise sources, that is, negative and positive points, the discharges from blunt points, and streamers and coronas.

Among other interesting facts discovered was that carrier-like bands of intensified noise existed at certain radio frequencies. Furthermore, it was observed that for a particular operating condition the frequency intervals between all these bands were more or less the same. Thus it appeared that radio interference was being produced by a regularly recurring phenomenon having a frequency of occurrence equal to the frequency difference between successive noise bands. For example, with a discharge of 52 μa from a negative needle point, strong interference was experienced at 2, 4, 6, 8, 10, and 12 mc, the noise voltage decreasing but the band width increasing with frequency. A linear curve could be drawn between discharge current and frequency in megabursts per second indicating the direct relationship.

Interposing a high-Q inductor at the end of a trailing wire antenna was suggested as a means of reducing static over a narrow band of frequencies. A resistor at the end of the trailed wire resulted in general noise reduction but not so much as was secured over the narrow band by the inductor.

7.1.3 Nonlinear Elements

The conclusion of the investigators was that the only hope of greatly improving radio reception through modification of the receiver circuits alone during periods of precipitation static lay in the use of nonlinear circuits or devices.

A Hallicrafter receiver with an audio limiter was found to be less susceptible to noise of the type examined than a receiver not so equipped. Analysis indicated that antenna limiters would probably not produce much benefit.

The fact that shock excitation produces long drawn-out oscillations in the various stages of a radio receiver led the investigators to believe that "dampers" of various sorts, which restrict the oscillations in the first i-f circuit before they have time to set the following circuits into shock oscillation, might prove to be very useful. Even though an improvement in a signal-to-noise ratio of only several times is obtainable by means of a damper alone, it is entirely possible that dampers used in conjunction with antenna, i-f and audio limiters, and other devices could effect a very material improvement in radio reception during adverse circumstances.

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[a] Project C-21, Contract OEMsr-92, Oregon State College.

7.2 STUDY OF EXISTING CHARGE DISSIPATORS

A study was made of several types of devices, each of which had as its function the discharge of accumulated charge on an airplane without at the same time producing radio interference. These devices were of self-ionized and pre-ionized types. The former were simply discharge wires or other structures in which the physical or mechanical construction was varied in the hope of attaining better discharge with less accompanying noise. Examples of pre-ionized types are radioactive cups and a flame discharger.

Point and fine-wire dischargers of several sorts were studied and compared to the conventional Bendix unit.

The conclusion reached was that the only kind capable of discharging large currents in a rapidly moving airstream with absolutely no measurable radio interference is the pre-ionized flame discharger. It consists of a high-temperature oxyacetylene flame at the end of a long, slender conducting electrode.

Another conclusion of the investigation was that the noise-current frequency spectrum obtaining quite generally for most gaseous discharges irrespective of pressure and polarity seems to be fairly constant up to roughly 5 mc and to vary inversely as the square of the frequency at higher frequencies.

Chapter 8

PRECIPITATION STATIC RESEARCH

Use of inverse vacuum-tube voltmeter for measuring high potentials and application to study of precipitation static.

8.1 THE PROBLEM

THE PURPOSE of this project[a] was to attack the precipitation static problem[1] by developing test equipment, making flights in storms, correlating the data secured and, if possible, develop means of eliminating the interference. Because a suitable airplane with which to collect systematic records under storm conditions was not available, the main part of the program could not be completed. The test equipment, however, was designed, built, and tested.

8.2 EXPERIMENTAL PROCEDURE

It was believed that an experimental investigation of the electric charges on and the electric potential gradients around airplanes flying through snow, rain, or sand storms would provide a basis for designing equipment to reduce communication failures in such storm conditions.

The first step in the investigation was the development of a generating voltmeter, together with suitable indicating and recording equipment, particularly adapted to the measurement of potential gradients at the surface of an airplane.

8.3 APPARATUS DEVELOPED

The apparatus developed consisted of wind-driven generating voltmeters which generated voltages proportional to the surface potential gradient at particular positions on the plane, vacuum-tube voltmeters for measuring the generated voltages, an instrument panel for mounting the microammeters for the several vacuum-tube voltmeters and standard aircraft instruments, a panel-mounted instrument for measuring vertical accelerations, and a motion-picture camera and drive for photographing the instrument panel.

The generating voltmeter was of conventional design and needs no description. After considerable experience with conventional vacuum-tube voltmeters with their positive-ion troubles, an inverse vacuum-tube voltmeter circuit was developed.

8.3.1 Inverse Vacuum-Tube Voltmeter

As is well known, the inverse vacuum-tube voltmeter employs a high-vacuum tube in a reverse connection, that is, the input signals are applied to the plate circuit, the grid current being a measure of the applied potentials. The plate is biased negatively so that positive voltages applied to the grid result in greater grid current flow. The circuit for the voltmeters finally employed is shown in Figure 1. The input impedance of this circuit (neglecting insulation resistance) is of the order of 5,000 megohms. The 9002 tube was chosen after some experimenting, although the tubes must be individually selected for the job. Another suitable tube of small size is the 6C4.

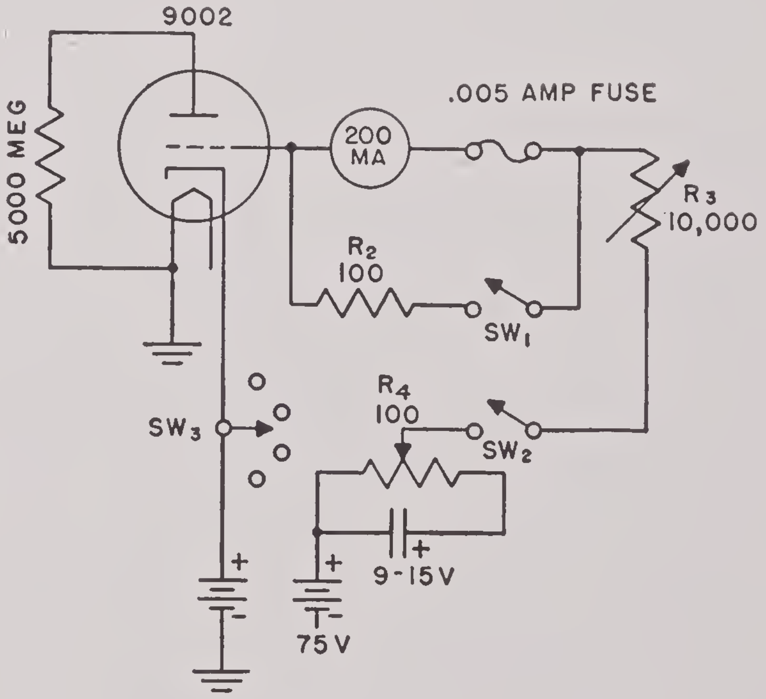

FIGURE 1. Circuit diagram of inverse vacuum-tube voltmeter.

In operation, SW_3 and R_4 are adjusted to give mid-scale reading on the microammeter with no voltages applied to the input. R_3 is used to adjust all tubes and instruments to the same sensitivity. The range of the instrument is changed by closing SW_3 and adjusting R_4 again to give the mid-scale reading. Calibration curves are shown in Figures 2 and 3.

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[a] Project C-41, Contract OEMsr-678, University of New Mexico.

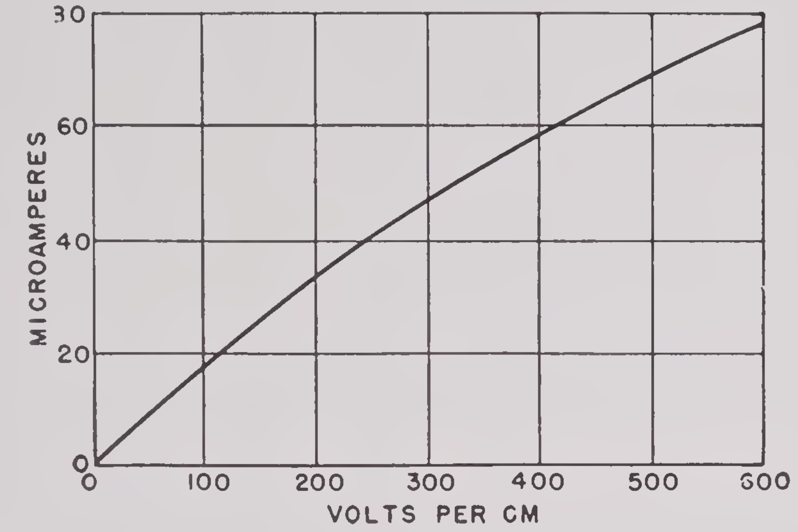

Figure 2. Calibration of voltmeter, low range.

8.3.2 Camera Drive

It was desired to photograph the entire instrument panel, containing five microammeters, air temperature, airspeed, and other essential data, at rates of 1 frame per second and 1 frame per 5 seconds. A Paillard Bolex 16-mm camera was employed, with a motor drive and gear box with shifting gears mounted directly to the small motor.

8.3.3 Other Equipment

The g-meter for measuring vertical acceleration consisted simply of a magnetically damped arm centered on two springs with a scale calibrated directly in g.

An all-metal plane capable of swift ascent and descent, equipped with two-way radio, de-icing equipment, blind-flying instruments, and oxygen for high-altitude work was desirable. A Consolidated B-24 stationed at the Alamogordo Air Base was made available and the instruments were installed. A short flight was made when weather conditions were appropriate and the plane returned to base. The total time devoted to gathering data by means of the plane amounted to 3 hours. The plane was then wrecked on a routine military mission not connected with the project.

The brief experience with the instruments and the plane indicated that the techniques developed were satisfactory and that if another plane could have been secured data of worthwhile nature could have been developed.

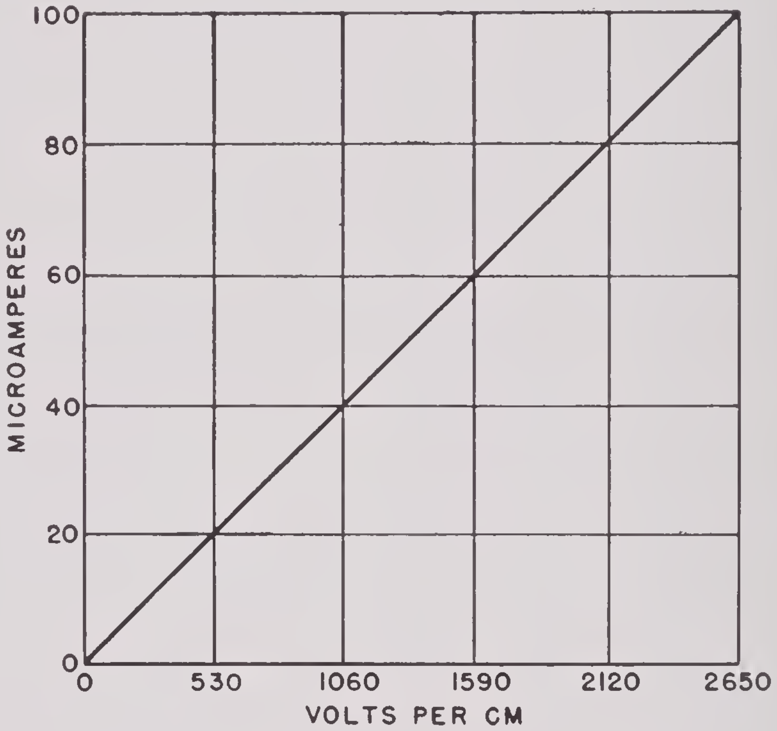

Figure 3. Calibration of voltmeter, high range.

Chapter 9

EFFECT OF AIRCRAFT SURFACE TREATMENT

Studies and tests yielding quantitative data upon the relationship between surface treatment of aircraft skin materials, specifically with military paint finishes, and the accumulation of electric charges known as precipitation static.

9.1 EQUIPMENT AND METHODS OF TEST

Equipment for blowing snow or frost crystals over sample surfaces and measuring the charging rate was developed during 1938 under a cooperative project between the Bendix Aviation Corporation and the Purdue Research Foundation. This equipment was used in the work done under Project C-64.[a]

The test equipment[1] consisted of a vacuum-cleaner motor and fan which supplied the air blast, a hopper with motor-driven screw-feed mechanism for feeding the snow into the air blast at a constant rate, an insulating support for the test sample, a flexible-tube connection for returning the snow to the upper part of the hopper where it is separated from the air stream by centrifugal action, and a portable galvanometer for measuring the charging current from the test sample to ground. The support for the test sample was semicircular in shape and made of paraffined maple 1½ in. thick. A recess $\frac{1}{16}$ in. deep and 1¼ in. wide was turned in the edge of the block to take the test samples, which were 1¼ in. wide and 12 in. long. The radius of the bottom of this recess was 3¾ in. In the center of this recess a rectangular groove 1 in. wide and ½ in. deep was turned to form a passageway for the snow-laden air along the inner surface of the sample. The test sample was held in place by an insulated wire spring which ran lengthwise along its outer surface.

Since the test sample was bent to form the arc of a circle having a radius of 3¾ in. and the snow-laden air was blown along its inner surface, the snow was held in contact with the entire length of the sample by centrifugal force.

The snow was fed into the air stream by the motor-driven screw feed at the rate of 30 cc of loose unpacked snow per second.

The velocity of the air stream along the sample was estimated by measuring the air speed in a straight extension of the rectangular curved passage by means of a standard Pitot tube and was found to be approximately 65 mph.

At an air speed of 65 mph and a feed rate of 30 cc per second, the average depth of snow on the sample was less than 0.05 mm, which permitted ample opportunity for the snow particles to come in contact with the test sample.

The galvanometer used to measure the charging rate was a Leeds & Northrup portable type having a deflection sensitivity of 57 divisions per μa. One terminal of the galvanometer was connected to the ground and to the metal part of the hopper containing the screw feed mechanism and to the fan and motor frames; the other was connected to the back side of the sample by means of a connection to the spring which holds the sample in place.

The Aircraft Radio Laboratory supplied samples of the following camouflage paints from which test specimens were prepared:

Dark O.D.	No. 41	Red	No. 45
Light O.D.	No. 42 (green)	White	No. 46
Gray	No. 43	Blue	No. 47
Black	No. 44	Yellow	No. 48

The test specimens were prepared by spray-painting a single coat of each paint on a carefully cleaned strip of Alclad skin-metal. To determine the effect of the metal under the paint, the Dark O.D. No. 41 paint was sprayed on aluminum, copper, Dowmetal F cleaned, Dowmetal F with the oxide left on, Duralumin and cold-rolled steel. The spray gun was carefully cleaned each time before changing to a different paint sample.

Later the Naval Research Laboratory supplied samples already coated on 1¼x12-in. strips of Duralumin. One sample of each of the following was provided.

Acetobutyrate, clear	2 coats
L-12 Gray	2 coats
M-485 Gray	2 coats
71 Line Gray	1 coat
Dope Red	2 coats

To provide information regarding the effect of temperature on the charging rate when using snow or frost crystals, it was proposed to make tests at approximately 20, 0, and —20 F. Since previous tests had shown that the same results were to be expected from the use of frost crystals from the refrigerator coils as from snow, frost crystals were used until snow became available.

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[a] Project C-64, Contract OEMsr-679, Purdue University. This work led to a much more extensive investigation for the Special Devices Branch, Aircraft Radio Laboratory, Wright Field, Contract W-33-038-ac-19 under the title *Precipitation Static Tests of Surface Coverings for Aircraft.*

9.2 SUMMARY OF RESULTS

The results of a large number of tests on camouflage paints and metals, with and without special surface treatment, have led to the following tentative conclusions:

1. A large percentage of the charging takes place at or near the point of initial impact of uncharged snow or ice crystals with the surface, indicating that the charging is due to the contact and separation of substances having dissimilar surface characteristics.
2. All of the camouflage paints became negatively charged under most conditions of test and usually charged at rates of 1.25 to 5 times that of aluminum. This indicates that they fall below aluminum in the triboelectric series.[2]
3. Metallic lead was the only material found which always became positively charged with snow and frost crystals, indicating that snow and frost crystals fall between lead and aluminum in the triboelectric series.
4. The charging rate obtained with snow varies widely, depending on the conditions under which the snow was formed and its subsequent history.
5. The charging rate of a paint appears to be affected very little by the metal under it.
6. In general the charging rates of the paints decreased with increase in temperature over the range from —10 to 20 F and fell to zero at some point between 20 and 32 F.
7. Dowmetal F that has just been cleaned with steel wool may charge either positively or negatively. However, if the surface is covered with a heavy coat of the gray oxide, it usually charges negatively at a low rate.
8. Oxidized lead initially charges negatively with frost crystals, but the oxide wears away rapidly and the charge reverses in sign.
9. The test results indicate that a large percentage of the charge acquired by an airplane flying through snow or ice crystals is generated on the frontal surfaces and on the propellers near the points of initial impact of the particles with the plane.
10. Under the predominant conditions where a plane flying through snow or ice crystals acquires a negative charge, it should be possible to neutralize largely the charging from the camouflage paints by metal-spraying a thin film of metallic lead on about half or two-thirds of the frontal surfaces, or by coating the propellers with metallic lead.

9.3 FURTHER INVESTIGATIONS DESIRABLE

The results of the work on this project indicate that the following further investigations are desirable.

1. Investigation should be made to find or develop a suitable binder for camouflage paints which would fall above aluminum in the triboelectric series when subjected to the same conditions of temperature and humidity that are encountered by a plane in flight.
2. A more extensive investigation should be made of the charging rate at various points along a surface corresponding to a wing section, in an effort to find means of causing the snow or ice particles to discharge back into the plane before being carried away by the air stream.
3. Since it is apparent from tests with segmented test samples that most of the charging occurs at or near the point of initial impact of the particles with the surface, the charging rate of the semiconducting rubber of the de-icers should be investigated.

Chapter 10

NOISE ELIMINATOR TESTS

Investigation of the manner in which an airplane collects a charge from precipitation, testing and improvement of the effectiveness as noise eliminators of several discharging systems, design and development of a simple form of voltage gradient indicator to be used as a lightning strike warning device.

10.1 INTRODUCTION

Although a number of the factors causing precipitation static and means for reducing the effect on radio communication systems were known before this project[a] was begun, one of the major accomplishments of the project was to collect in one report data collected during actual flight in precipitation static areas.

In brief, the project[1] proved that the use of high-voltage or high-frequency devices as a means for aiding in discharging the plane while in flight would be no more effective than an unenergized point discharger. The flight tests proved that high gradient fields can be detected and their relative position and strength indicated during flight with simple measuring instruments and exploratory prods. One of the conclusions reached during the investigation was that a painted surface of an airplane collects charge at a higher rate than a clean surface. This fact was substantiated by other investigators and was instrumental in influencing the Military to remove paint from all noncombatant airplanes. (See Chapter 9.)

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[a] Project C5-68, Contract OEMsr-893, United Air Lines.

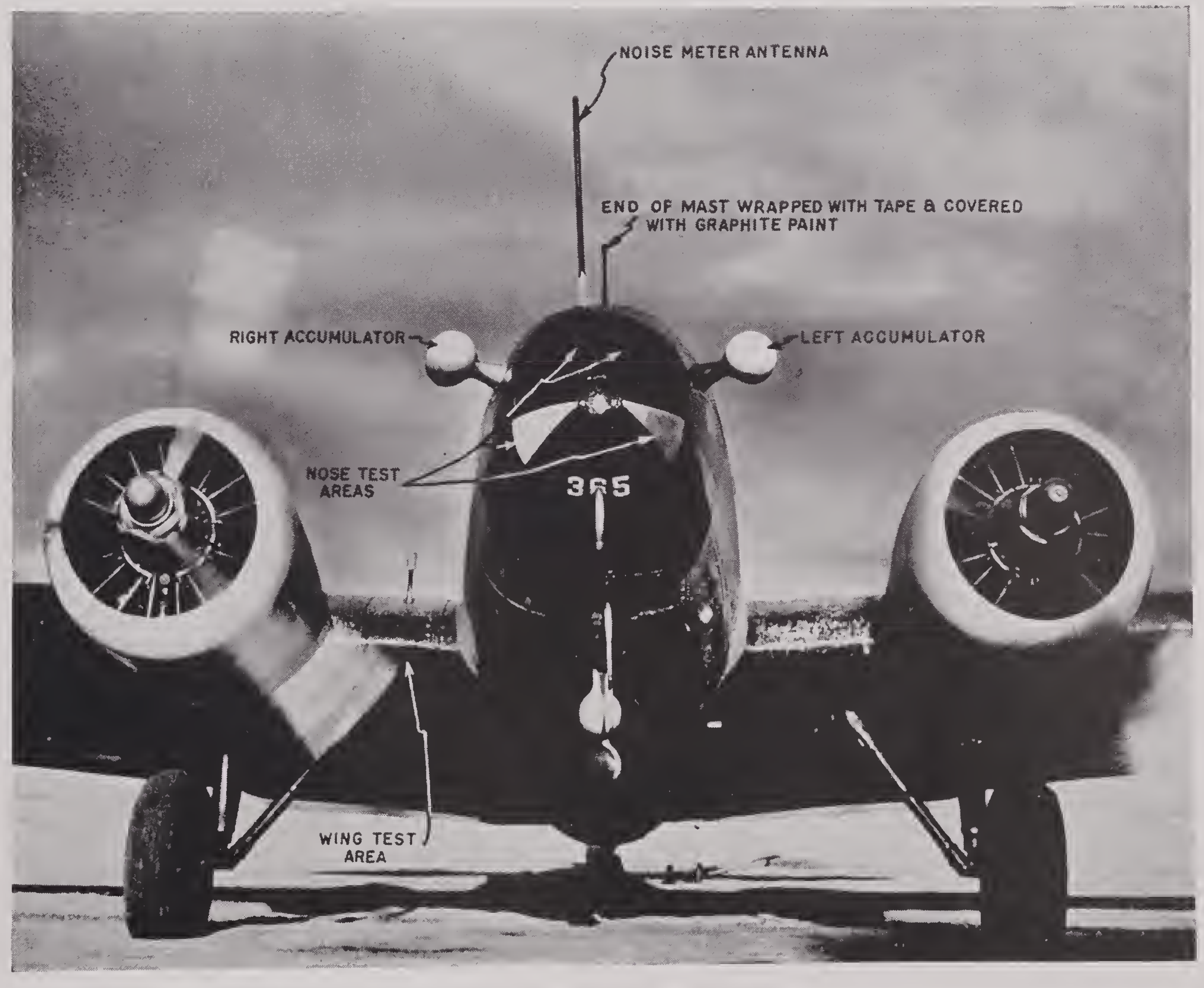

Figure 1. Boeing 247-D all-metal plane used in precipitation static research.

10.2 APPARATUS EMPLOYED

A major portion of the experimental work was conducted in a Boeing 247-D all-metal plane (Figure 1). Some auxiliary testing was carried out with a Douglas C-47 military cargo airplane and other tests were made in a Consolidated C-87 cargo version of the Liberator bomber.

Test areas insulated from the airplane were faired into the surface so they would not interfere with the normal flow of air. These test areas were intended to show how charge is accumulated and what portion of the plan structure is responsible for the greatest accumulation. The propellers were equipped with full-length de-icer shoes and fine bare copper wires were cemented in the leading edge of these rubber shoes. A slip ring was mounted on the propeller hub so that connections could be made to these fine wires. A second slip ring was connected to the propeller itself.

Spurious corona discharges from uncontrolled points about the airplane were believed to be responsible for much of the radio noise. To prevent this uncontrolled factor from entering the results of the investigations on the several discharge devices, all unshielded and sharp projections about the airplane were covered in some manner, usually by a generous application of rubber tape covered with a coating of conducting graphite paint. Antenna insulators, masts, and tension springs were all treated in this fashion. Filters were placed in the antennas so that the static current discharged by them could be measured without interrupting radio communication.

Trailing wire dischargers were fitted to the wing tips, stabilizer tips, and tail position. These wires were retractable.

Gradient indicators consisting of small point-terminated dipole explorers were located parallel to the major axis of the plane so that components of the electric field could be measured in all directions simultaneously. The generating voltmeter discussed in Chapter 8, consisting of a rotating vane and amplifier, was mounted on an inspection door on the under surface of the right wing. The location of the search equipment is shown in Figure 2.

10.3 INSTRUMENTATION

Several methods of measuring the small currents encountered in precipitation static were employed. In one case, an RCA 991 neon tube shunted by a capacitor flashed at a rate proportional to the current flowing (Figure 3C). This method was simple

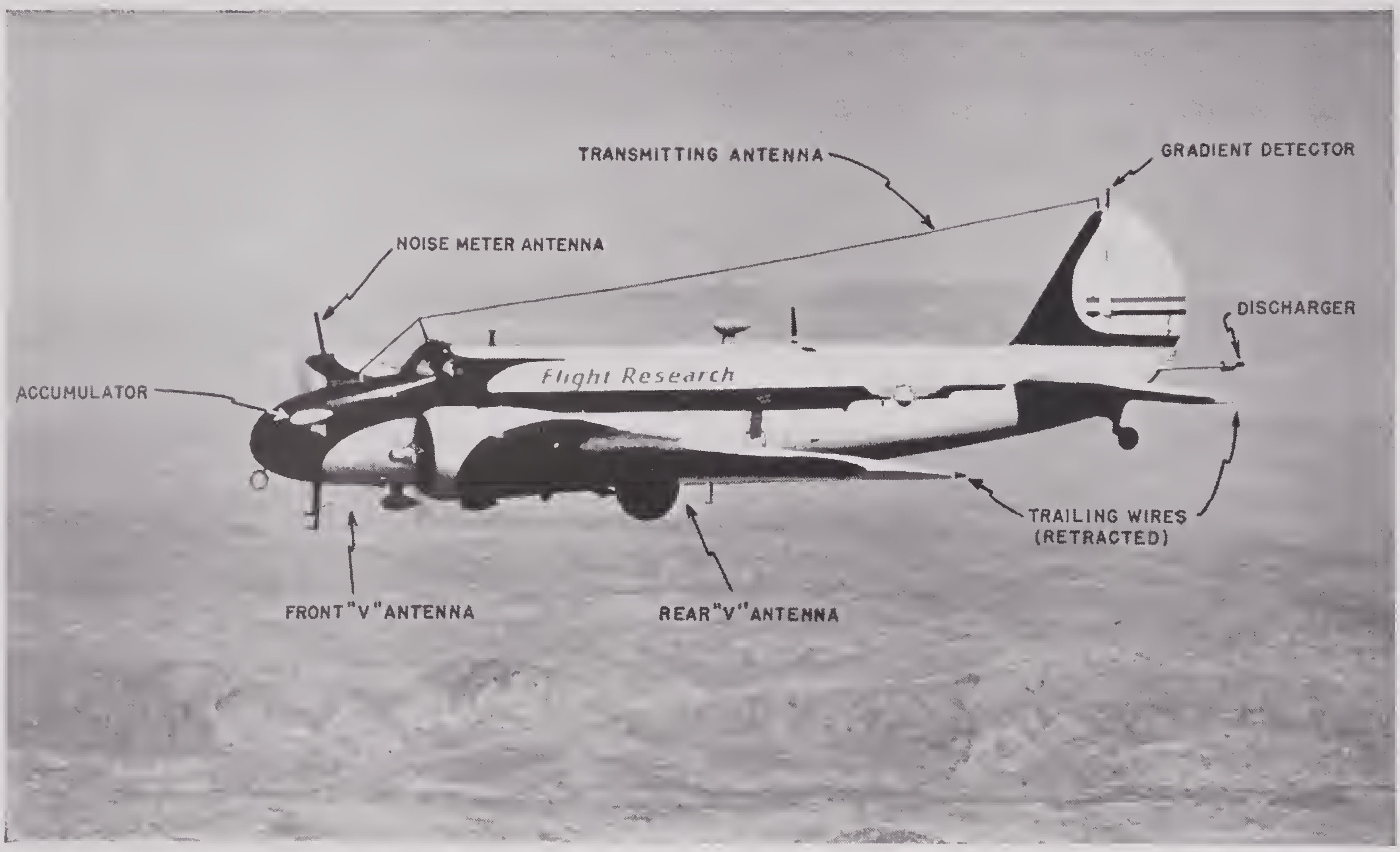

Figure 2. Location of research equipment external to cabin.

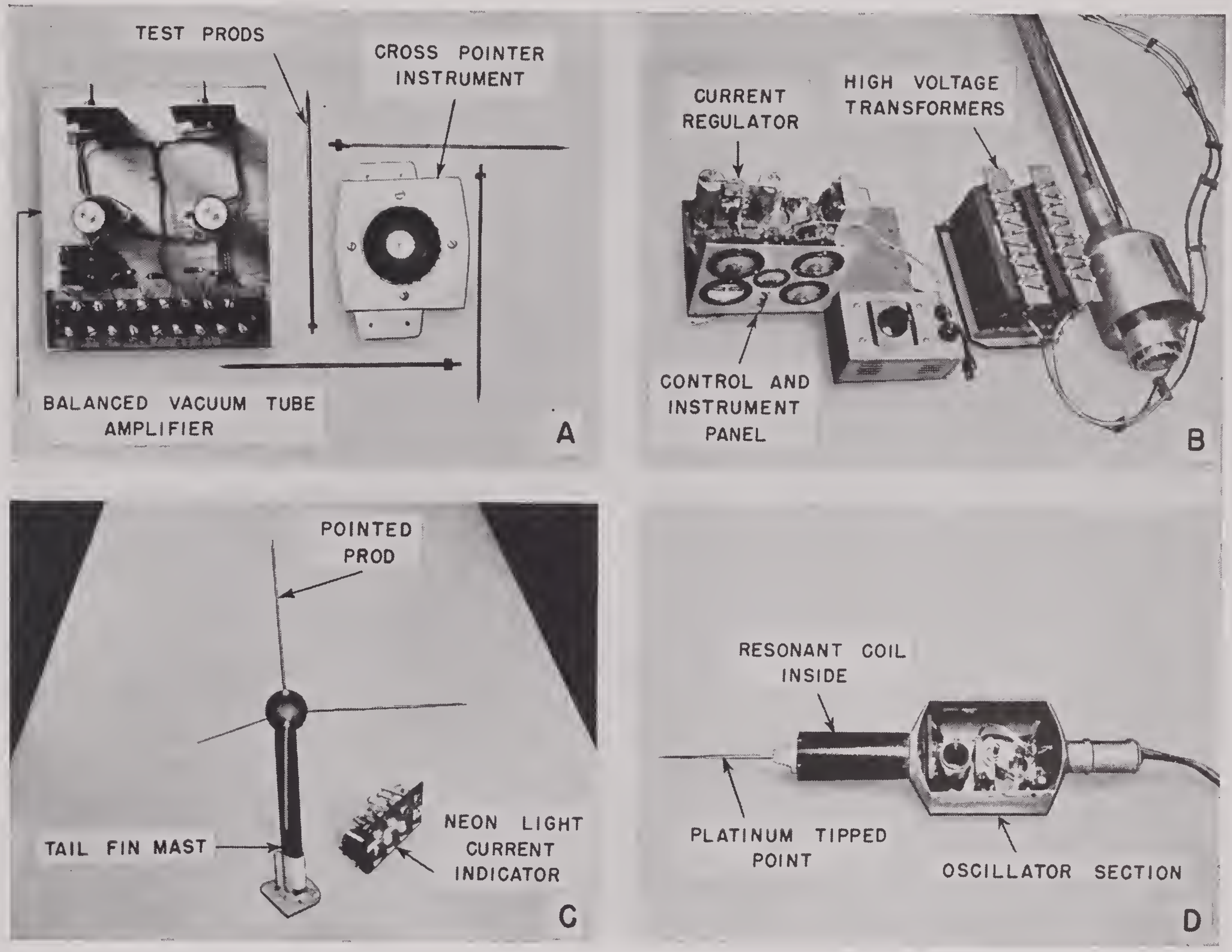

Figure 3. Electric and electronic measuring instruments employed in United Air Lines research.

and economical but proved unreliable at currents of ½ μa or less. In other cases a single-stage amplifier with a plate milliammeter (vacuum-tube voltmeter) was found to be quite successful. Currents of 0.05 μa per scale division were easily measured. Twenty-four of these amplifiers were built on a single chassis.

10.3.1 Charge Indicator

A basic requirement for the analysis of data obtained was a knowledge of the static potentialities of the various weather areas encountered. For this purpose a charge indicator was constructed and used on all flights. This instrument consisted of an accumulator element streamlined in form and insulated from the plane. It collected its charge from precipitation in the same manner as did the plane. That is, certain portions were subject to impact by snow, ice, or rain; other portions were subject to frictional effects, and other portions of the surface were subject to charge by separation. Charge rates were measured by noting the current flowing from collector to the airplane. Once the airplane had acquired a high potential with respect to space, discharge would occur, reducing the plane potential with respect to the accumulator element so that the charge on this element would flow to the airplane.

Provisions were made so that the information on the several current meters could be photographed simultaneously by means of a single-frame 16-mm camera.

10.4 TEST RESULTS

Studies were made of several types of dischargers, all compared to a trailing wire consisting of a 5-ft length of very fine wire connected to a suppressor resistor made of approximately 5 ft of rubber-covered Aquadag-impregnated rope, the resistance of which was from 250,000 to 500,000 ohms. This trailing wire was definitely helpful in reducing static noise. Under severe conditions, currents as high as 200 to 400 μa were observed, but in each case the trailing wire lowered the noise level appreciably.

Other trailing wires of various types, such as those with terminals made up of a brass sphere or a cluster of three very sharp points, were used, but none seemed to have any advantages over the simple wire described above.

A device of the Slayter Electronics Corporation consisting of a high-voltage supply and a discharge element made up of sharp points or long conducting rods was tested in flight. Both units were intended to produce ionization about the discharge points and thus to aid in discharging the plane.

In test flights, it was demonstrated that the Slayter apparatus "produced no improvement in radio reception, while the trailing wire discharger rendered the receiving equipment usable although all the static was not removed."[1]

After many flights with several models of dischargers, including wick devices and those which are energized by high-voltage 60-cycle or r-f power, the conclusion was reached that random noises must be controlled by shielding to reduce the potential gradients at the points producing the noises; in this manner more current would be discharged by the controlled discharging system which must be capable of discharging current in the order of 200 to 300 μa.

Flight tests showed that the majority of the charge accumulated by an airplane comes from the frontal surfaces. The amount of accumulation is a function of speed and surface coating. A painted surface will collect greater charges than will an unpainted surface.

Test flights indicated that the gradient indicator, consisting of prods attached to a vacuum-tube voltmeter, would indicate the presence of thunderclouds when they were several miles distant. In general, an instrument capable of registering 5 to 10 μa would mark the existence of a thundercloud five miles away, this current increasing to 30 to 50 μa when the plane approached within a half mile of the cloud.

Using four prods, one each placed in wing tips, tail, and nose parallel to the longitudinal and lateral axis of the plane, and with the indicator taking the form of a cross-pointer instrument (Figure 3A) marked to show the direction toward the danger area and the safe limit of field intensity to avoid danger of lightning strike, warning of a possible lightning strike was found to be distinctly feasible.

Chapter 11

THE BLOCK-AND-SQUIRTER SYSTEM

Study of the order of magnitude of corona discharge currents from large military aircraft; flight tests to investigate the efficacy of various discharge systems; development of the "block-and-squirter" system in which the plane is discharged in pulses between which the radio receiver is operative, the system relying on the phenomenon of persistence of hearing by which an operator takes no notice of short intervals of silence in reception.

11.1 INTRODUCTION

Precipitation static radio interference was considered to be one of the most serious hazards encountered in the training of Army pilots on large military aircraft. For this reason, the Commanding Officer of the Second Air Force assigned one of his staff planes, a B-24 Liberator four-engined bomber, and later a B-17 Flying Fortress, for use by Washington State College [WSC] in conducting research studies of precipitation static.[a]

Prior to undertaking this research with the Second Air Force and with OSRD, WSC had done considerable research over a period of some five years in the field of radio static interference on commercial aircraft. With the cooperation of United Air Lines, studies had been made of receiver performance and of balancing networks for reducing static shocks. Therefore the work on this project[1] was, in effect, a continuation and enlargement of the work already underway.

11.2 CAUSES OF PRECIPITATION STATIC

The most serious type of aircraft radio interference occurs when the plane flies through precipitation, such as rain, snow, or frost crystals, or through dust-laden air. This kind of interference is especially serious because it occurs when radio aid is most needed, that is, during times of limited or zero visibility.

Various theories have been advanced as to the mechanics by which an electric charge is generated on a plane in flight, and how it is collected and discharged. The cause of radio interference from precipitation static is generally understood to be due to corona discharges taking place between various parts of the plane and air. These commonly take place from sharp metal corners and edges of wings and ailerons, from antenna wires, or from any sharp or pointed metal part of the plane exposed to the outside air, and probably also from the sharp edges of propellers.

There appear to be two general conditions under which a plane accumulates an electrostatic charge which causes radio interference. In the first case, the plane flies through a highly charged area such as a thundercloud in which there exists a high space gradient in potential, and the metal plane seems to act more or less as a conductor between differently charged areas. In such areas, measured charging currents on the moving plane usually change rapidly in value and polarity. In the second case, the plane appears to collect a potential charge at a more or less uniform rate and to discharge it at about the same rate. Under these conditions, the plane is usually negative with respect to any discharge which takes place. This second condition prevails more frequently than the first and therefore is considered the more serious hazard to flying, although both conditions create radio interference in the receiver on the plane.

A secondary source of radio interference arises because the propeller surfaces engage the air at much higher speed than do the other surfaces of the airplane so that whether the accumulation of an electrostatic charge is due to impact, friction, or sweeping action, it is probable that a potential difference exists between the propeller and the plane. Interference from this current flow through the propeller bearings would closely resemble "wheel static" in automobiles and could be corrected by providing an electrically conducting path around the oil film. Hydraulically operated propellers have no ground brush but electrically operated propellers do have a good ground brush and would not cause this type of interference.

Radio interference occurs on all frequencies commonly used on the plane, although not always to the same degree. However, under severe static conditions all radio contact is interrupted. It is agreed that the radio-beacon frequencies, 200 to 400 kc, are the most important to the pilot needing radio aid, and therefore all observations during these tests were made on this band.

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[a] Project C-70, Contract No. OEMsr-848, Washington State College.

11.3 PURPOSE OF THE PROJECT

The expressed purpose of Project C-70 was to make flight tests of precipitation-static radio-interference phenomena, and of devices for their control. First, it was desired to measure the magnitude and characteristics of corona discharge currents which cause radio interference on large military aircraft. Before intelligent steps can be taken to suppress, neutralize, or otherwise control radio interference on these planes, it is necessary to have more information than was available on the subject. Measurements on four-engined bombers had not been undertaken prior to this contract.

Second, it was desired to flight-test such devices as might be developed by WSC or by other contractors and having for their purpose the suppression or control of precipitation static interference on large aircraft. Facilities were especially favorable for such tests because of prior arrangements with the Second Air Force where planes and experienced instrument pilots were made available for flying in inclement weather.

The work naturally fell into three divisions: development of equipment for making the desired measurements in flight, organizing and conducting test flights in weather favorable to static interference, and carrying on supplemental laboratory tests of equipment and devices which had promise of reducing the interference.

11.4 FLIGHT TESTS

Flights were made to determine the magnitude of discharge currents which take place on large military aircraft, to ascertain the rate at which a charge accumulates on a plane, and to test devices developed for producing a noiseless discharge, or for suppression or control of the discharge. Most of the flights were made in the Pacific Northwest, although some thirty-three states were crossed during the course of the work. Test data were secured at altitudes ranging from 3,500 to 32,000 ft, over mountains as well as plains, in rain, dust, snow storms, and through frost crystals, at temperatures from +20 C to below —40 C.

11.5 TEST EQUIPMENT

The essential measuring equipment consisted of multiband radio receivers, a radio noise meter, d-c microampere amplifiers, and Esterline-Angus recording milliammeters. A standard 6K7 amplifier tube operated as a d-c amplifier increased the small currents obtained from the test devices to a value where milliamperes could be recorded on the recording milliammeter.

To measure the rate at which the plane accumulated an electrostatic charge, a standard radio-compass loop housing was copper-plated and mounted on a spar in free air in front of the nose of the plane. This charge collector, having a projected area in the line of flight of exactly 0.45 sq ft, was grounded to the plane through a 15-megohm resistor. The current through this resistor was measured by a d-c amplifier and represented 45 per cent of the rate per sq ft at which the wings, elevators, etc., collect electrostatic charges. (See Figure 1.)

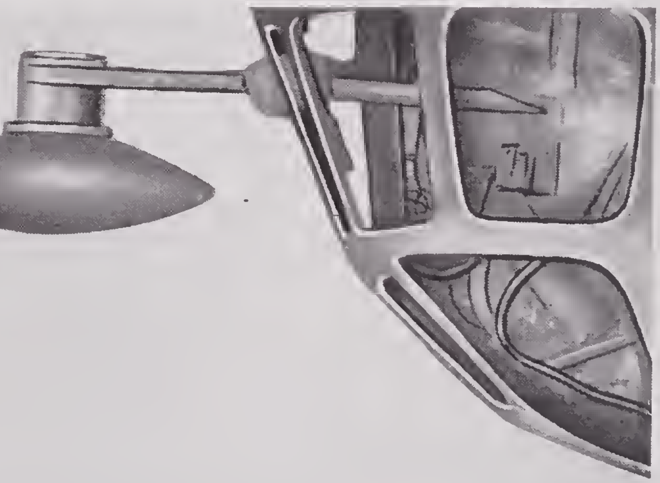

Figure 1. Copper-plated teardrop automatic direction finder housing mounted on wooden spar and adjusted in position in front of nose of plane. Wire through center of spar connects teardrop to d-c amplifier to measure rate of accumulation of charge on plane.

Continuous records of all of the test instruments were made during flight. One recording meter registered the charging rate, another the radio signal and interference noise, and the other two of the four meters were employed to test the several types of discharge devices tested. The final report[1] includes 220 sheets of flight data collected by means of this equipment. The data include discharge rates obtained from Bendix wires, trailing antennas, rod dischargers, bristle devices, etc. In all, 23 different types of discharge systems were examined.

11.6 STUDY AND ANALYSIS OF FLIGHT TEST DATA

1. Many flight tests were made on B-17's and B-24's in rain, snow, and frost crystals, and discharge cur-

rents were measured. These range from the usual 25 to 100 μa up to 500 or 600 μa in more severe cases.

2. When flying through a local thundershower, discharge currents were measured in excess of 2,500 μa. Polarity of the charge on the plane reversed when flying through a thundercloud.

3. The polarity of the plane was usually negative with respect to the discharge wires extended from the tail of the plane.

4. Radio noise usually started when the discharge current on the Bendix wire reached 35 to 100 μa.

5. The most severe radio noise was always encountered when flying through frost crystals.

6. The use of multiple discharge wires did not assure noiseless radio reception at high discharge currents.

7. The application of high-voltage alternating current to accelerate corona discharge did not reduce static interference.

8. The application of high voltage or high frequency to the receiving antenna did not aid radio reception.

9. An increase in discharge current accompanies an increase in airspeed of the plane.

10. The cadence in static noise ("wow" static) noted on some of these flights appears to be associated with the difference in speed of adjacent propellers.

11. The trailing antenna wire will discharge current approximately in proportion to its length and is no less noisy than a Bendix wire.

12. The discharge characteristics of the trailing antenna can be improved by adding a tuft of small wire bristles to the antenna weight.

13. A 100-ft Bendix wire will discharge approximately in proportion to its length and was not found to be any less noisy than a 10-ft Bendix wire.

The upshot of all this work indicated that:

1. The magnitude of corona discharge current on a four-engined bomber appears to be too large to be dissipated noiselessly and at the same time keep uncontrolled discharges from taking place elsewhere on the plane.

2. Higher-speed planes with larger frontal areas will undoubtedly experience even larger corona discharges than take place on the B-17's and B-24's.

3. Properly applied a-c potentials to a suitable discharger should keep the charge potential on the plane low enough to prevent uncontrolled discharges from taking place.

11.7 THE BLOCK-AND-SQUIRTER SYSTEM

Attempts to produce noiseless discharge as a means of solving the static problem having failed, attention was turned to another approach to the problem. Since persistence of hearing, like persistence of vision, takes no notice of small intervals of silence in reception, it was thought that if the discharge on the plane could be alternately blocked and accelerated and if the receiver could be alternately opened and blocked in synchronism, then radio reception might be possible during the worst kinds of static troubles.

The method of putting such a method into operation is as follows:

A suitable a-c potential is applied to a corona discharger system on the plane, depleting the electrostatic charge sufficiently during one half cycle so that during at least part of the opposite half, when the applied potential blocks the discharge current, the effective corona voltage on the entire plane will be low enough so that no corona discharge will take place from any other part of the plane.

During the quiet, or blocking, part of the cycle the radio receiver is turned on and accepts a clear signal. During the squirter, or noisy, part of the cycle, the radio receiver is turned off and hence it is not cognizant of the severe noise accompanying the discharge of the plane.

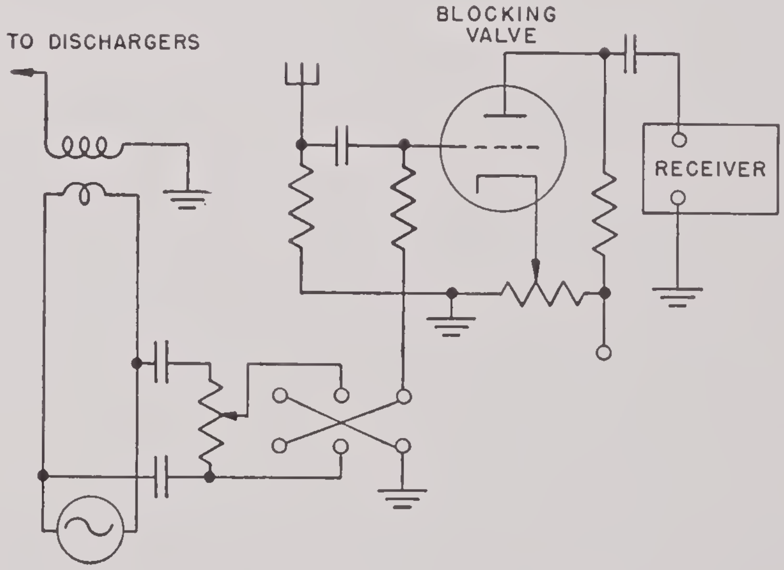

Figure 2. Pulsing and blocking, or "static squirter" system.

The receiver valve is interconnected with the a-c generator supplying the squirter system and properly phased therewith so that the receiver is always turned off during the noise discharge. (See Figure 2.) The frequency of the a-c generator should be above the audio range of the receiver, perhaps between 10 and 15 kc.

The metal plane constitutes an effective electrostatic capacitance which various authorities have estimated at from 500 to 2,000 $\mu\mu f$. The electrostatic charge is accumulated at a measurable rate while flying through precipitation, the rate depending upon conditions.

For a given set of conditions, the corona discharge from some point on the surface of a plane in flight will follow the equation:

$$I = KE(E - E_s)$$

in which I is current, E is the total effective voltage applied to cause I, and E_s is the potential at which corona starts. K depends upon the geometry of the discharge point, and upon various other conditions, such as air velocity and pressure, humidity, etc. Experimental curves derived from tests on a Bendix wire closely follow this equation. It is evident then that with a properly designed corona discharger system and with all sharp points and corners removed from the plane, the E_s of the discharger system can be kept considerably below the E_s of any other part of the plane.

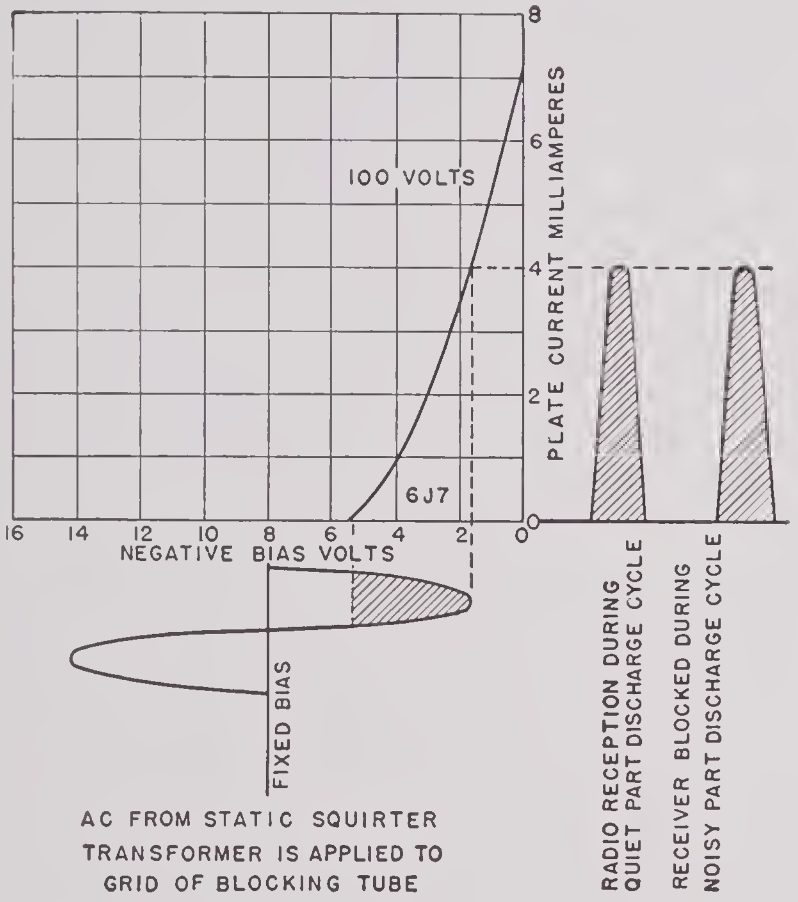

FIGURE 3. Typical performance curve of sharp cutoff tube over-biased to be used as blocking valve.

If the effective d-c potential E could be varied slowly up or down from some certain value, the discharge current would increase or decrease according to the above equation. Practically, this effective corona potential E may be varied by superimposing an a-c potential and the discharge current thus alternately increased and decreased.

By means of laboratory tests it was proved that, if a high enough a-c potential is superimposed on the d-c corona, the discharge can be reduced during part of the cycle to a noiseless value, or even completely stopped. On the other half of the cycle, the discharge will be greatly increased. This ejects the electrons from the discharger in a succession of squirts.

Theoretical analysis and some laboratory tests indicated that the effective corona potential for a discharge of 100 μa from a Bendix wire would not be less than 45 to 50 kv; that the a-c power required for the block-and-squirter system would be of the order of 10 watts or less; that the a-c potential should have a frequency of the order of 10 to 15 kc.

11.8 FINAL CONCLUSIONS

It was concluded at the close of the contract that radio interference on large military aircraft is caused by such large corona currents that a successful noiseless discharge system would not be practical and that a system based on the principle of discharging the plane in a series of pulses between which the receiver is turned on offers considerable promise of success.

The block-and-squirter system was further developed and flight-tested with the Second Air Force and a prototype unit built by General Electric Company for use on B-29's. The original test equipment weighed about 750 lb. The prototype weighs less than 40 lb and occupies a very small space in the dorsal fin of the plane. V-J Day came before the prototype was available for installation on the B-29's, but the Army approved continuation of the tests.

PART IV

PANORAMIC RECEPTION

Several projects in Division 13 were concerned with panoramic receivers, either for monitoring purposes or for more active participation in communication taking place in war zones, i.e., for jamming. The following projects dealt with receivers per se:

C-27, a moving-screen receiver for the region 500 to 600 mc.

C-36, an improved receiver for the band 3 to 10 mc, also a similar improved receiver for the band 0.1 to 30 mc. As part of the project, a report, *Fundamentals of Panoramic Reception,* was issued.

C-39, a scanning and stopping receiver for the 350- to 750-mc range.

Uses of panoramic principles for interference generation (jamming) are covered in Part V.

In the summaries of this work to follow, the chronological order in which the work was done has not been followed. Instead, the portion of the C-36 final report dealing with fundamentals is reviewed first, then follow the reports on the several receiver projects. The summaries of reports on the applications of panoramic principles have been confined, largely, to abbreviated descriptions of the apparatus developed. Where significant statements or quantitative data contained in the final reports of the receiver projects relate to principles of panoramic reception, they have been lifted from the report in which they appear and included in Chapter 12.

Chapter 12

THE FUNDAMENTALS OF PANORAMIC RECEPTION

A summary of the parts played in panoramic reception by such factors as proper frequency allocations, design of scanning filters on resolution and signal-to-noise ratio, types of indicators, determining frequency of received signals, recording signals, reception of pulse signals, receivers without frequency sweep, uses of panoramic receivers, and other factors of importance to the general subject.

12.1 INTRODUCTION

THIS REPORT, based to a large extent upon work carried out in certain NDRC projects,[1,2,3] summarizes some of the fundamentals of the art of panoramic radio reception.[a] By panoramic reception is meant the reception of signals present within a band of frequencies and the display on a visual indicator of information concerning these signals. The received signals are usually radio signals but may be other types. Devices in which a variable oscillator is employed to obtain a visual diagram or record of the frequency characteristic of an equipment unit or circuit component will not be included under the term panoramic receivers.

There are two general methods of panoramic reception: (1) that in which the signal frequencies are swept in succession past a relatively narrow selecting circuit, the output of which is used to actuate the visual indicator, and (2) that in which the signal frequencies are applied simultaneously to a number of selecting circuits of adjacent frequencies, and the outputs of these circuits are applied in rapid succession to the visual indicator. In this report principal consideration will be given to the former method, only a brief section being devoted to the latter. A theoretical alternative to (1) is to move the selecting circuit past the signals, but the practical utility of this appears to be very slight except possibly for special problems.

The process of moving the frequencies past a selecting circuit is commonly termed scanning. It is accomplished by automatically varying the frequency of a beating oscillator, known as a sweep oscillator. The essential elements of a panoramic receiver of the frequency-scanning type are shown in Figure 1. Other functions frequently provided in a panoramic receiver, in addition to those of scanning and visual indication, are aural reception and precise frequency determination.

The process of determining the different frequencies present in a given band by heterodyning the signals in succession past a sharp selecting circuit is quite old in the art.[4,5] Almost as old is the provision of automatic sweep for the beating oscillator.[6] Panoramic reception adds to this frequency-analyzer technique the simultaneous display on a visual indicator of the different signals in the band which is being scanned.[7,8]

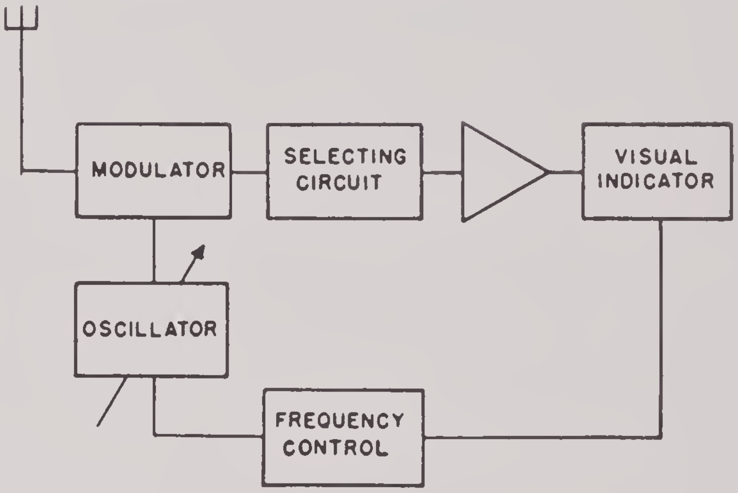

FIGURE 1. Essential elements of panoramic receiver.

Panoramic reception has much in common with ordinary radio reception. Thus in each case (a) provision must usually be made for handling a wide range of signal strengths; (b) high sensitivity is desirable, the limiting factor being random electric disturbances in the medium or the equipment (conveniently referred to, even in a visual device, as noise); (c) heterodyne technique is advantageous to obtain the desired gain and selectivity.

Much more important than the similarities, however, are the differences between the two types of reception. Some of these are:

1. Whereas the ear is effectively able to appreciate only one signal at a time, the eye can receive at almost the same time information concerning a number of signals.

2. In panoramic reception it is necessary to swing one or more beating oscillators continuously, and fairly rapidly, over a band of frequencies, whereas in ordinary radio reception the beating oscillator frequency is adjusted manually and remains fixed during a desired response.

3. Unless the circuits which precede the scanning of the radio band are tuned in synchronism with this

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[a] Project C-36, Contract No. OEMsr-357, Bell Telephone Laboratories, Inc., Western Electric Co., Inc.

scanning, these circuits must have a reasonably flat response over a frequency band as wide as the scanning band. This is important in the design of antennas, transmission lines, and r-f amplifiers.

4. A wide range of signal intensities presents greater problems in panoramic reception.

5. Because of the finite build-up and decay times of a selecting circuit, the width of the selecting circuit desirable for use in the scanning process depends upon the rate at which the frequency range is scanned. Hence the width of the receiver band is usually considerably greater for a panoramic receiver than for an ordinary receiver.

6. As a result of the various factors noted above, the selectivity or resolution obtainable in a panoramic receiver is poorer.

7. The signal-to-noise ratio and useful sensitivity are poorer than can be attained in an ordinary receiver.

8. In panoramic reception there are many different ways of presenting the information, and different types of indicators may serve different purposes.

9. The characteristics of vision must be taken into account in panoramic reception.

12.2 FREQUENCY ALLOCATIONS

Choice of a frequency allocation for a panoramic receiver, that is, choosing of the frequencies for beating oscillators, i-f circuits, etc., is apt to prove difficult. Factors entering into this choice are (1) the location and width of the frequency band or bands to be scanned, (2) the capabilities of sweep oscillators, (3) the avoidance of multiple responses, and (4) the selectivity realizable at different frequencies.

12.2.1 Scanning Band

The frequency band to be scanned may be located almost anywhere in the frequency spectrum. The width of the scanning band is ordinarily determined either by the fineness of resolution desired or by the number of signals that can be simultaneously monitored.

Frequently when a panoramic receiver is to cover a wide frequency band, it may also be desired to scan alternatively narrower bands capable of being positioned anywhere within the wider band. This is because a narrowing down of the band provides better resolution and fewer signals to claim the operator's attention. However, either breadth or variety in the scanning band is apt to be reflected in a multiplication of i-f circuits and complexity of equipment.

12.2.2 Sweep Oscillators

Sweep oscillators used in panoramic reception are mostly of two types:

1. Electronically controlled oscillators, that is, f-m oscillators whose frequency is varied by means of a reactance tube which applies a quadrature voltage to the oscillating circuit. The amplitude of the quadrature voltage is ordinarily determined by means of a sawtooth wave applied to the reactance tube so that each sweep of the frequency band is followed by a rapid return.[9] With such an oscillator, the location and width of the scanning band may be readily changed by simple potentiometer controls acting upon the sawtooth wave. Using conventional oscillator circuits, electronic sweep oscillators can be built for frequencies up to approximately 100 mc and reasonable linearity can be obtained for a sweep of the order of 20 per cent of the top frequency.

2. Mechanically controlled oscillators, that is, oscillators whose frequency is controlled by using mechanical motion to change capacitance, inductance, or both, in the oscillator circuit. The element which is varied may be a capacitor, a Lecher wire, or a cavity. Normally with this form of oscillator the sweep through the frequency band in one direction and the return in the opposite direction are at about the same rate. Since it is difficult to obtain coincidence of the traces for the two directions of scanning, blanking of one scan may be necessary. Mechanically controlled oscillators are particularly useful in the frequency range above that of electronically controlled oscillators. The percentage of frequency swing may be quite large.

Other ways of controlling the frequency of an oscillator are possible, but have not been used to any substantial extent. One of these, for example, is to control the oscillator inductance by varying the d-c flux through the core of an h-f inductance coil.

By means of one or more frequency multipliers the frequency band of a sweep oscillator may be moved upward, the percentage frequency band remaining unchanged. Suitable suppression of unwanted frequencies is necessary. In case the percentage frequency band covered by a sweep oscillator is less than that desired, it is usually possible to obtain a greater percentage by building the sweep oscillator for a higher frequency. The sweep frequencies may then be modulated down to a lower location if desired. Inversely, the percentage sweep necessary to scan a given frequency band may be reduced by shifting the signal band upward in frequency.

If two or more different widths of scanning band are desired, and if the ratio of the maximum scanning band to the minimum scanning band is quite large, it may be difficult to provide both wide and narrow sweeps in the same oscillator. This is because the frequency control for the narrow sweep, whether electrical or mechanical, becomes too fine. A practical solution of this problem may be to provide more than one scanning oscillator and a corresponding number of i-f amplifiers. Thus, a first oscillator might scan the entire band and a second oscillator a part of the band. All scanning oscillators must, of course, be controlled by the same sweep circuit. The different scanning oscillators may be used alternatively or, if desired, combinations may be used for wide-band scanning. Multiple scanning oscillators furnish an effective way of reducing the ratio of top frequency to scanning band for any individual oscillator but they tend to complicate filtering and shielding problems. All circuits which precede any scanning oscillator must be as wide as the scanning band of that particular oscillator.

12.2.3 Heterodyne Methods—Multiple Responses

Heterodyne technique is desirable to obtain the selectivity and gain for panoramic reception. It is well known that one of the most important problems in using the heterodyne method is the avoidance of multiple responses. Let us consider first a single heterodyne receiver, one in which a single intermediate frequency is employed. The most important type of unwanted response is the so-called image or second channel response which occurs when the intermediate frequency is less than half the width of the band to be covered.[10] If the frequency band to be covered has a ratio of top to bottom frequency of more than 3 to 1, there is no intermediate frequency below the band for which image response will not occur. In addition to image response, there are a number of other types of response which may be serious, including higher-order combinations of signal and oscillator frequencies. If the intermediate frequency of a single heterodyne receiver is above the band to be scanned, the selecting circuit employed in the scanning process must be at a relatively high frequency and it is usually impractical to obtain the desired selectivity characteristics.

Image response, as well as other undesired responses, can be reduced by using r-f tuning, controlled either electronically or mechanically so as to track with the beating oscillator. This, however, is difficult in a wide-band panoramic receiver.

A way of avoiding the multiple response difficulties of single heterodyne reception is to employ instead a double superheterodyne method, with the first intermediate frequency placed above the signal band. This method, however, leads to other possibilities of multiple response which must be taken into account.[11] The principal trouble in this case is due to spurious responses resulting from higher-order combinations of the two beating oscillators. In practice it has been found difficult, even with extremely careful shielding, to reduce this type of response to a satisfactory value unless the frequency allocation is such that only difference frequencies resulting from the fourth or higher-order harmonics of the two oscillators can yield the first intermediate frequency. This requirement means that the first intermediate frequency must be more than three times the highest frequency in the signal band, the precise ratio depending upon the selectivity available in the two i-f circuits and the frequency of the second i-f stage. Also the frequency of the first beating oscillator should be placed above the first intermediate frequency.

Even when these conditions are fulfilled, careful design is necessary to minimize different types of undesired response. Also there are two rules to be observed. First, to avoid spurious responses due to insufficient shielding or selectivity, it is desirable that no intermediate frequency or beating-oscillator frequency should fall within the input band. Second, to avoid double response the location of the intermediate frequency should not be less than half of the width of the preceding i-f circuit as measured between attenuating regions. This means that it is impracticable to go from a very high intermediate frequency to a very low one. Thus if the i-f amplifier operates at approximately 1,000 mc, and if the 3-to-1 ratio mentioned above is preserved, an upper frequency of about 300 mc for the signal band is indicated.

12.2.4 Selectivity

The characteristics of available selecting circuits play a large part in determining the location and number of i-f stages in a panoramic receiver. There are several reasons for this. First, the requirements for the final selecting circuit, which determines the receiver resolution, are much more rigorous than those for the final selecting circuit in an ordinary receiver. This exerts an important influence on the choice of the last intermediate frequency. Secondly, rapidity of attenuation rise in any i-f circuit affects the location of the succeeding intermediate frequency as noted above. Another rule is that the pass bands of all preceding circuits should be wider than the final selecting circuit so as not to affect the performace of that circuit.

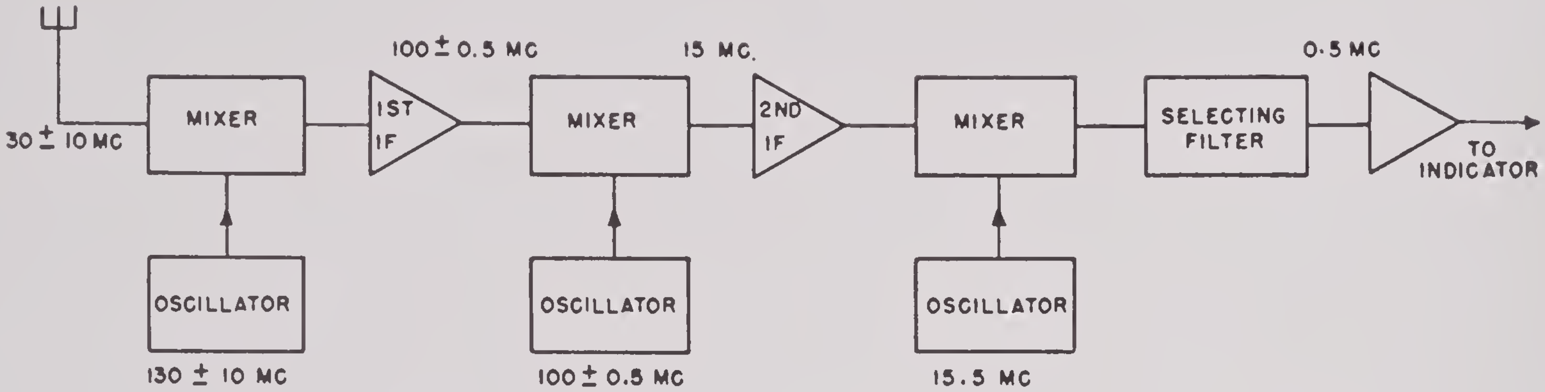

FIGURE 2. Illustrative frequency allocation plan.

Limitations of realizable selecting circuits therefore tend in the same direction as both sweep oscillator limitations and considerations of multiple response, namely, toward additional numbers of i-f stages. Hence triple or quadruple heterodyne technique may be desirable for panoramic reception.

To illustrate the frequency allocation principles discussed above, there is shown in Figure 2 a possible allocation diagram for a panoramic receiver intended to scan either the complete frequency range of 20 to 40 mc, or, alternatively, a band of 1 mc anywhere in that range. Triple modulation is employed. To simplify the sweep oscillator design, two different sweep oscillators are provided, one for the wide-range sweep and one for the 1-mc sweep. At any one time, one of these would be automatically varied while the other would remain fixed. The third beating oscillator is invariable.

The first intermediate frequency is placed at 100 mc, or slightly more than three times the top frequency of the input band. The pass band of this i-f stage must be as wide as that of the succeeding sweep oscillator, i.e., 1 mc. The second intermediate frequency must be higher than half the width of the first i-f circuit between cutoff points but should preferably lie below the input band. Accordingly it is placed at 15 mc. The pass band of this second i-f circuit must be wide enough not to affect the scanning filter characteristic. The final intermediate frequency is placed at 500 kc, which is a suitable but not necessarily preferred location for the selecting filter.

The allocation shown in Figure 2 is merely one out of a practically infinite number which would accomplish the same result. For a wise choice, careful study of the capabilities of components and comparison of at least a moderate number of different allocations would be essential.

12.3 AUTOMATIC CONTROL OF SIGNAL INTENSITY

In panoramic reception, the input signals may range from one or more microvolts up to several millivolts, an intensity range (also referred to as volume range) of 60 db or more. The maximum intensity range which can be applied to types of indicators commonly used is from 5 to 20 db, depending on various factors. Accordingly it is necessary to provide some method of automatically reducing the signal intensity range ahead of the indicator. This requirement can be met by a device which acts upon the signals after they have been selected in the scanning process.

A more fundamental and more serious disadvantage of a wide intensity range is that a strong signal cannot be cut off sufficiently rapidly in the final selecting circuit to prevent some masking of a nearby weak signal, the result of which is a loss of resolution. As discussed below, no satisfactory way of getting around this difficulty is available.

Devices for reducing the range of signal intensities have been used in the radio and telephone arts for many years.[12,13] Similar devices may be used to reduce the intensity range in panoramic receivers but the performance requirements are somewhat different. Devices useful for panoramic reception are of two general types (see Figure 3). First, there is limiting or gain control arranged to maintain substantially constant output over a certain range of input and usually having constant gain below this range. Second, there is the compressor, whose input-output characteristic when plotted on a db scale is a straight line with a slope less than unity. With either device a threshold may be provided to suppress the output when the input is below a certain minimum value. Either one would consist of a circuit (sometimes referred to as a vario-losser) in which the loss or gain is changed in accordance with the amplitude of the signals, either directly or by means of an auxiliary control circuit.

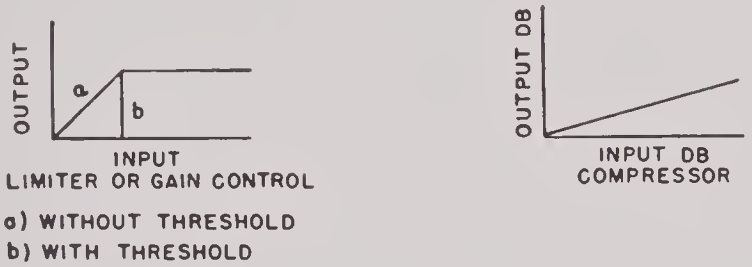

Figure 3. Characteristics of devices for intensity control.

Since there exists, prior to the final selection, no basis for obtaining a differential action as between signals, the intelligence necessary to operate the device for controlling intensity range must be obtained after the desired signal has been selected from other signals. Moreover, any control device introduced ahead of the final selection would probably produce objectionable intermodulation between signals. A possible way of reducing the signal intensity range ahead of the selecting process is to employ tuned rejection circuits to cut down signals that are especially high. Such an arrangement, however, involves a sacrifice in resolution, and becomes difficult and expensive where a number of stations with large signal strengths occur in the frequency band.

The wanted signal delivered by the final selecting circuit consists of brief spurts of energy when signals are encountered. The intensity control device may operate either from the a-c signal or from the corresponding rectified signal. Distortion is of no consequence, since the only use made of the high-frequency pulses delivered by the selecting circuit is to rectify them to obtain a visual indication. Owing to the brief duration of each signal, it is generally desirable that the action of the device used to reduce the intensity range should be practically instantaneous both in attack and release. This makes it important to consider time constants carefully and may rule out devices in which a backward-acting circuit controls the vario-losser.

While either limiting or compression may be used, compression is usually preferable, since it preserves some distinction between signals of different amplitudes. One advantage of such a distinction is that it facilitates differentiation between a weak signal and the transient response associated with a much stronger signal. A compressor in the form of a cascade limiter, comprising a series of interstages arranged so that successive interstages produce a limiting effect for increasing values of input, has been found useful.[2]

12.4 DESIGN AND PERFORMANCE OF SCANNING FILTER

The circuit which selects in succession the different signals in the frequency band may be a single filter or tuned circuit, or it may consist of one or more tuned amplifier stages. In any case, it will be referred to below as the scanning filter. The resolution of a scanning receiver, that is, the minimum frequency separation for which it is possible to differentiate between two signals of adjacent frequency, is dependent upon the design of this filter. Assuming a given filter design, the resolution obtainable depends upon the level difference between input signals. A further factor in resolution is the capability of the indicator. However, it will be assumed in this section that the indicator is able to utilize the resolution obtainable with the scanning filter. The subject of indicator resolution will be taken up subsequently.

When signals having broad spectra are to be resolved, as for f-m transmission or transmission of short pulses, the design of the scanning filter and the resolution which it affords become of much less importance, since these types of signals are placed farther apart in frequency and since the boundaries of any one signal spectrum will be less clearly defined. The following discussion therefore refers largely to signals of fairly narrow spectra.

12.4.1 Outline of Problems

Because filters are made up of inductive and capacitive elements which have appreciable time constants, and because the input frequencies are swept past the scanning filter, the design and performance of this filter must be considered in terms of transient response. The principles involved therefore differ from those for a filter or selective circuit for ordinary reception.

The signals applied to the scanning filter are in the form of a succession of frequency-modulated waves whose rate of frequency change is the scanning speed. Assuming negligible retrace time in the frequency sweep, the scanning speed is equal to the product of the frequency band swept over and the repetition rate. When the scanning speed in radians per second per second is large in comparison with the time constant of the scanning filter, then the filter response is stretched out, which produces confusion between signals of adjacent frequency and hence brings about loss of resolution. If, on the other hand, the rate of frequency change is very small compared to the filter time constant, then the filter is wider than it needs to be and the resolution consequently poorer.

The width of the scanning filter and the scanning speed affect also the signal-to-noise ratio for the scanning receiver. In particular there exists for any given scanning speed a value of filter band width which yields optimum signal-to-noise ratio.

Problems to be considered in connection with the scanning filter therefore include, first, determining the optimum design of scanning filter from the standpoint of (a) resolution and (b) signal-to-noise ratio; and second, the related but nevertheless distinct problem of determining the performance obtained with a given design of filter under different operating conditions. While the basic theoretical approach to these problems is in terms of the transient response of the filter, the results can be expressed in terms of its steady-state response characteristics.

The complete problem of scanning filter design is extremely complex. It was first studied a number of years ago.[14] Recently it has been treated more comprehensively, but so many variables are involved that solutions thus far have had to be based on simplified and somewhat idealized assumptions. However, theoretical work has been confirmed and supplemented by experimental results which have been particularly useful in furnishing quantitative relationships.

12.4.2 Analogy with Quasi-Stationary Filter Response

A qualitative picture of the effect of scanning filter characteristics on resolution may be had by analogy with the more familiar problem of the response of a filter to an a-c pulse the fundamental frequency of which is the mid-frequency of the filter. Assume that such an a-c pulse is applied to a filter which is critically damped, i.e., whose characteristics are such that the output rises quickly to its steady-state value but does not exceed it. Assume further that the filter has a discrimination of 12 db or more per octave of sideband frequency. It can be shown that for such a filter the build-up time approximates $1/B$ where B is the band width of the filter as measured between 6-db points. For this case, when the length of applied pulse is equal to or greater than the build-up time of the filter, then the duration of the response as measured between the half-amplitude points is the same as the length of applied pulse. If, however, the length of the applied pulse is less than the build-up time, the duration of response is equal to the build-up time, so that the response is stretched out and reduced in amplitude.

Now consider what happens in scanning. As an input frequency is swept past the scanning filter, the output response takes the form of an a-c pulse. The rectified response corresponding to this pulse is somewhat similar to that obtained by turning on a mid-frequency wave when the scanned signal reaches the leading edge of the filter and turning this off as the scanned signal passes the trailing edge. The nominal duration of this assumed output pulse is

$$T_A = \frac{B}{\gamma}, \qquad (1)$$

where B = width of scanning filter in cycles per second and γ = scanning speed in cycles per second ($\gamma = nF$ where n = repetition rate and F = frequency band swept over). The band width B is measured between points where the filter discrimination is 6 db.

If the actual pulse is assumed to correspond to a mid-frequency pulse of duration T_A, then the simple theory outlined above furnishes a rough criterion for determining the duration of the response. The duration of response when multiplied by the scanning speed gives the apparent band width of response, which may be taken as an approximate indication of the resolution obtainable between signals of equal amplitude.

If the resolution is plotted against the filter band width B for a fixed scanning speed γ, a minimum occurs when the filter is just wide enough to transmit the nominal pulse of equation (1) without any substantial decrease of amplitude and stretching out of response. This minimum determines the optimum band width for the assumed scanning speed.

Consider next the law of variation of optimum band width with change of scanning speed γ. Assume that the band width is left fixed and that γ is increased above the optimum for that filter. This stretches out the response and decreases the pulse amplitude. To restore optimum design, it is evidently necessary to increase the band width. However, such increase gives a double benefit in that the length of pulse to be handled increases at the same rate as the ability of the filter to handle a given pulse length. The change in band width necessary to restore optimum design corresponds to the change in the square root of scanning speed. That is,

$$B_o = K_B\sqrt{\gamma}, \qquad (2)$$

where B_o = optimum width of scanning filter in cycles per second and K_B may be termed the bandwidth factor. As before, filter band width is measured between 6-db points.

The minimum or best value of resolution occurs when the filter band width is about equal to $\sqrt{\gamma}$ ($K_B = 1$), the exact value depending on the filter damping. This value of scanning speed ($\gamma = B^2$) may be termed the critical scanning speed.

If now the filter band width B is left fixed while the scanning speed γ is varied, a theoretical curve can be obtained for resolution as a function of scanning speed. Below the critical scanning speed the resolution remains practically constant, while above the critical value the resolution becomes poorer directly as scanning speed increases.

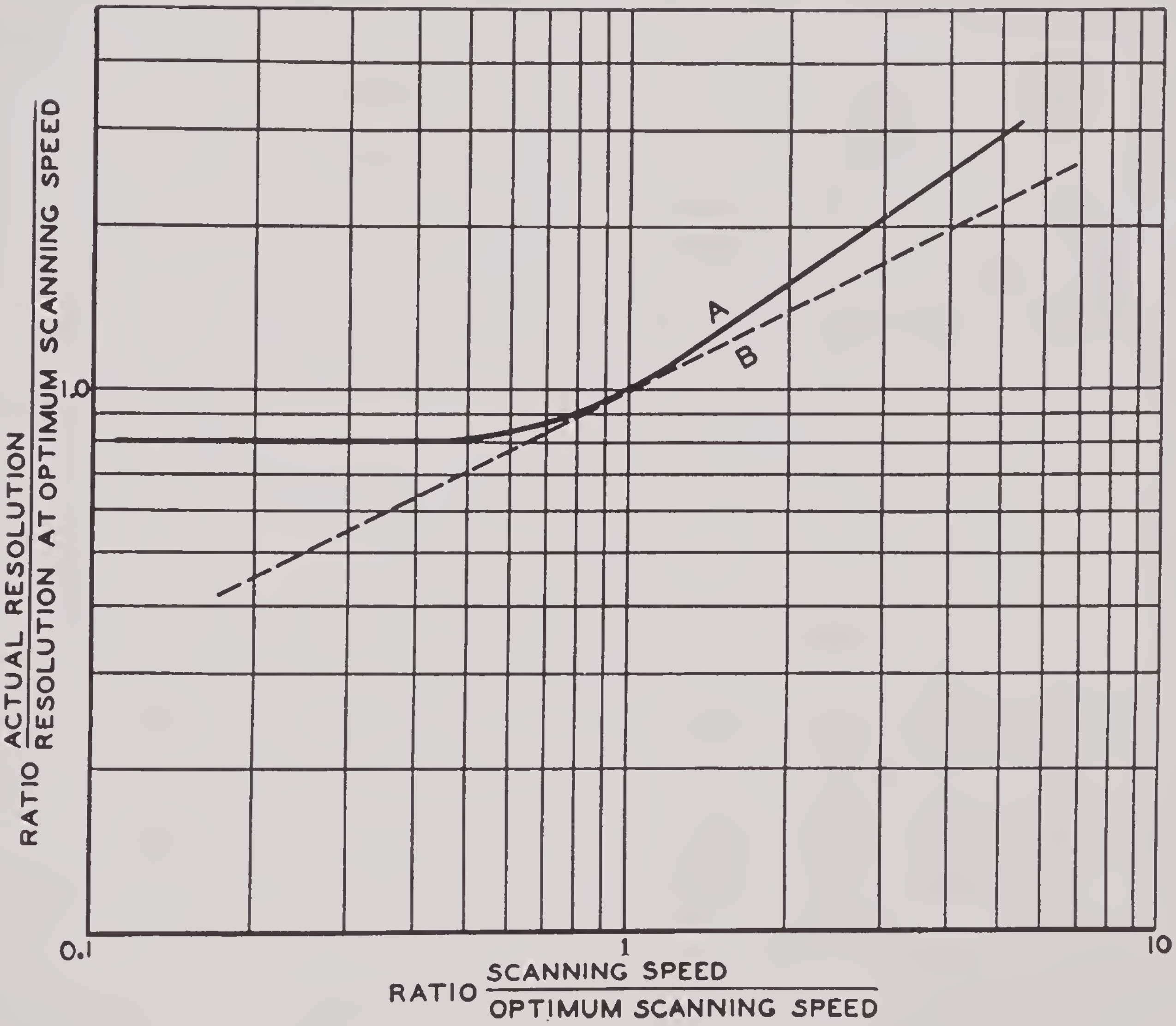

Figure 4. Resolution versus scanning speed.

12.4.3 Experimental Values of Band Width and Resolution for Equal-Level Signals

The foregoing discussion is based on simplified assumptions which account inadequately for the transient phenomena associated with scanning. Experience with a number of actual filters has indicated how the simple theoretical relationships must be modified for practical use.

Equation (2) for optimum filter band width has been found to apply reasonably well. However, the value of K_B varies considerably, depending on the filter design. Since it is usually not convenient to vary the filter band width, the scanning speed may be varied instead and a curve obtained from which the value of K_B may be computed. Values between 0.8 and 2 have been observed.

The optimum value of resolution S_o for a given filter may be indicated by the following equation:

$$S_o = \frac{K_S}{K_B} B_o = K_S\sqrt{\gamma}. \qquad (3)$$

Values of the resolution S for a given filter may be determined by varying the scanning speed. A typical curve of S versus γ, using equal-level signals, is shown at A in Figure 4. Above the critical scanning speed the resolution becomes much poorer, approaching direct proportionality to scanning speed, while below the critical speed it is more or less constant. This curve is intended for purposes of illustration and not for precise computation. More detailed experimental results are presented elsewhere.[1] Curve B shows the locus of best resolution when the filter band width is made optimum for each scanning speed. The point where the two curves are tangent represents the optimum or critical scanning speed for the assumed filter band width.

From such determination of resolution the value of K_S in equation (3) may be determined. This value varies considerably, depending upon the type of filter, type of indicator, and the level difference between signals being resolved. With equal-level signals and with an indicator affording maximum resolution, the spread of observed values of K_S is from about 1.2 to 3. The value of 1.2 was obtained for a filter which also had a value of $K_B = 1.2$, this being the best combination thus far obtained.

A chart showing optimum filter band width or optimum resolution as a function of scanning speed, assuming K_B or $K_S = 1.2$, appears in Figure 5. Filter band width or resolution for other values of K_B or K_S may be readily obtained by multiplying the values shown on the chart. A nomographic chart may be employed to derive optimum filter band width or optimum resolution directly from the three variables: (1) scanning band, (2) repetition rate, and (3) K_B or K_S.

12.4.4 Effect of Level Difference

Thus far consideration has been limited to the case where the two signals to be resolved are of equal level. For this condition the filter may be designed with a band width determined by assuming an appropriate value of K_B and with reasonably flat delay between the 6-db points. The amount of cutoff required beyond the 6-db points will depend upon the type of indicator, but in general a moderate cutoff, of the order of 12 db or more per octave of side-band frequency, would be satisfactory.

If a large level difference exists between the two adjacent signals, then the performance of the filter in the cutoff region assumes greater importance. This is for the reason that if a strong signal is not cut off rapidly enough it will mask a nearby signal, producing loss of resolution. Delay distortion in the region adjacent to the pass band of the filter may also give

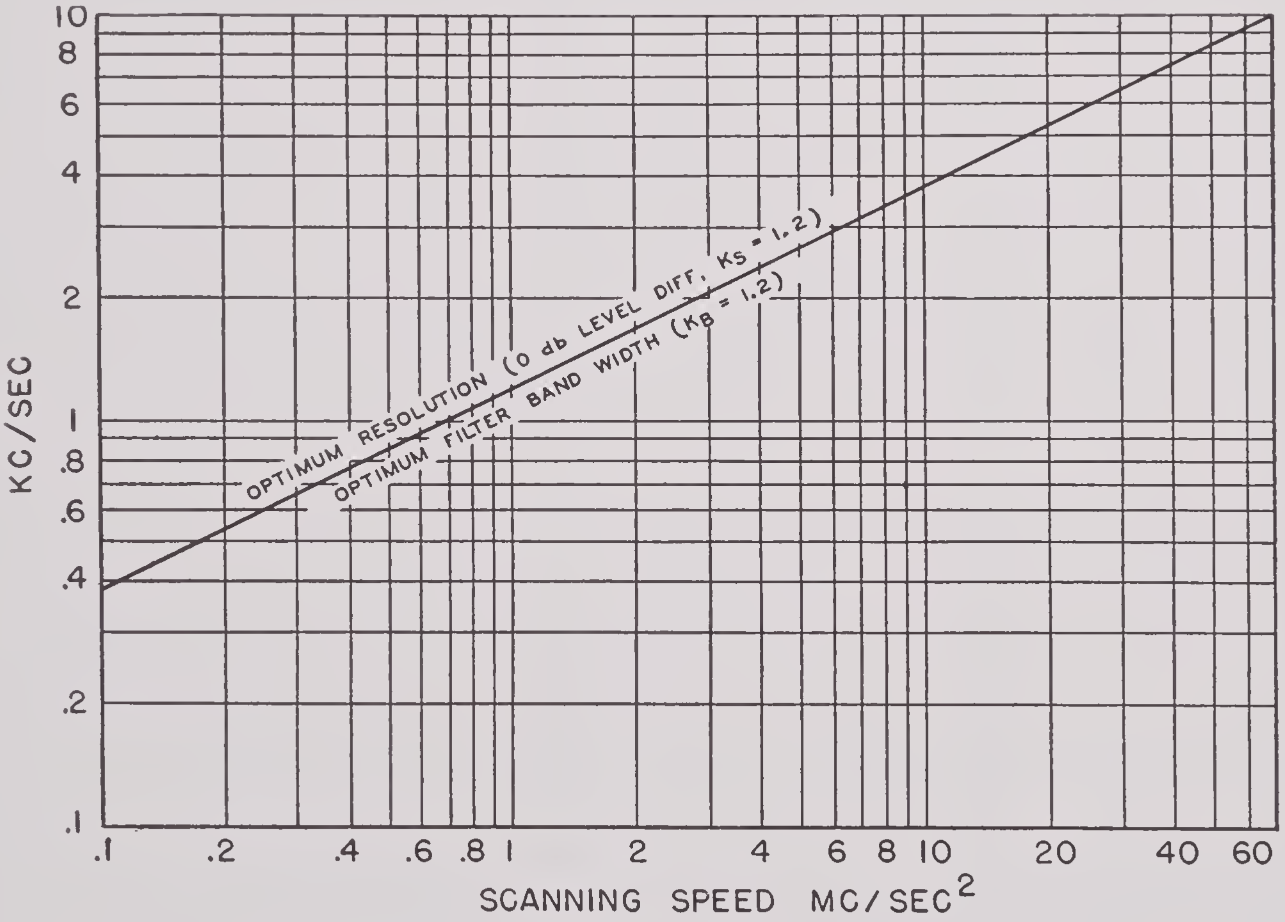

Figure 5. Chart for determining filter band width and resolution as a function of scanning speed.

rise to substantial impairment of resolution. These effects are most pronounced when the larger signal is scanned first, so that the trailing edge of the response is unduly prolonged. Hence, in designing a filter to be used for large level difference, it is necessary to obtain rapid and sustained cutoff together with reasonably small delay distortion extending as far into the cutoff region as practicable. A cutoff which becomes asymptotic to from 18 to 40 db per octave of side-band frequency, depending upon the level difference, is desirable.

Observations of resolution indicate that the optimum scanning speed for any given filter design is about the same for relative signal strengths ranging from 0 to 40 db, so that the filter band width information already presented for equal-level signals may be used in the case of unequal levels.

With the best present technique, the value of K_S increases with level difference somewhat as follows:

Level difference (db)	K_S
0	1.2
20	2.9
40	5.3
50	12.0

With poorer filter design or poorer indicator resolution the values of K_S for large level differences are much higher.

12.4.5 Signal-to-Noise Ratio

It has been pointed out that the band width required for the scanning filter tends to make the signal-to-noise ratio for a scanning receiver much poorer than that obtainable in a nonscanning receiver. Here it is interesting to consider the relation between scanning filter band width and signal-to-noise ratio. If the filter band width is chosen for best resolution, then the noise power varies inversely as the square root of the scanning speed. Hence, as low a scanning speed as practicable for best resolution is advantageous for signal-to-noise ratio as well. Looking at the filter purely from the standpoint of signal-to-noise ratio, two questions are of interest: (1) What is the optimum band width for a given scanning speed? (2) With a given band width, how does the signal-to-noise ratio vary as the scanning speed is changed?

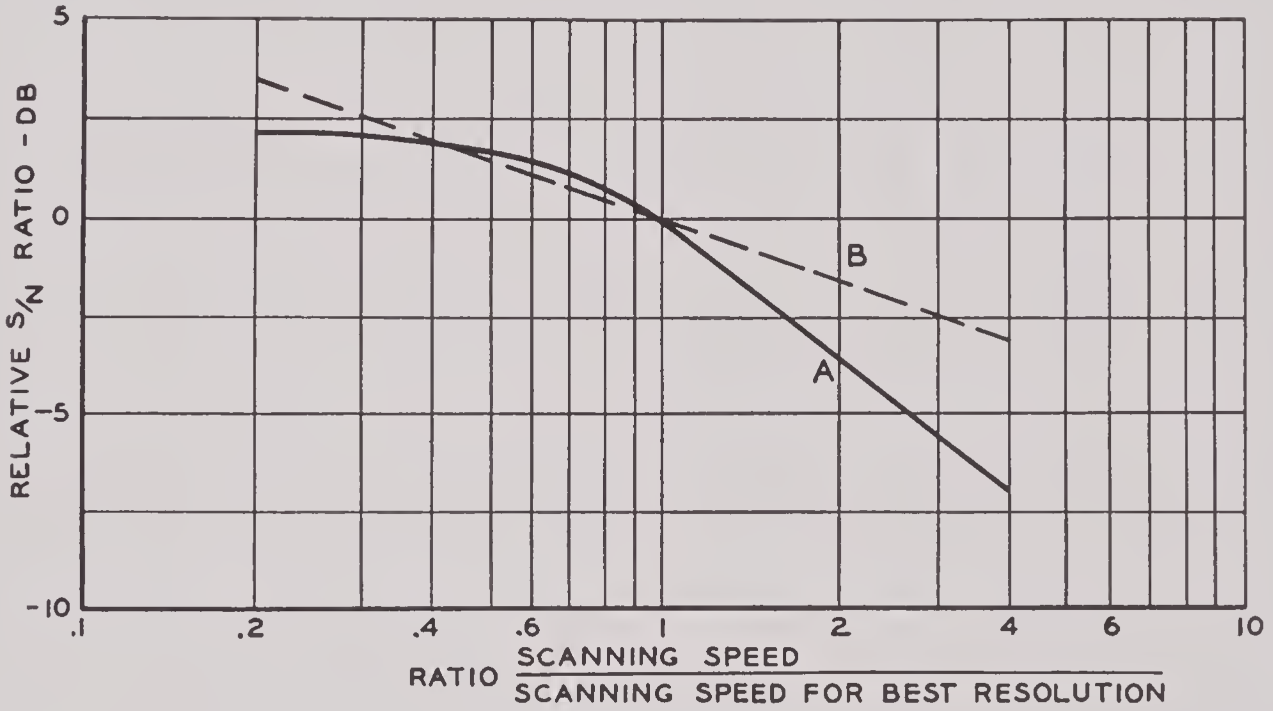

FIGURE 6. Signal-to-noise ratio versus scanning speed.

Experiments using a fixed filter have shown that the noise remains unchanged as the scanning velocity is varied over a wide range. However, as the scanning speed is increased above the critical resolution speed, the signal response decreases, so that the signal-to-noise ratio is reduced. Below the critical speed the signal response remains substantially constant, so that the signal-to-noise ratio is also constant. In this range, however, it would be possible, by using a narrower filter, to obtain substantially the same signal response with less noise. The complete curve of relative signal-to-noise ratio versus relative scanning speed therefore takes the general shape shown in curve A of Figure 6, while curve B shows the locus of signal-to-noise ratios when the filter band width is optimum for each scanning speed. The difference indicates the impairment in signal-to-noise ratio incurred by using a filter at a scanning speed other than that for best resolution.

On the basis of experiments along this line, it appears that the best signal-to-noise ratio for a good scanning filter occurs at a scanning speed quite near the optimum for good resolution.

12.4.6 Use of One Filter at Two or More Scanning Speeds

Under certain conditions it may be desirable, in the interest of apparatus economy, to operate a single scanning filter at two or more different scanning speeds. For such conditions the type of results to be expected as regards resolution and signal-to-noise ratio are indicated in Figures 4 and 6. These curves should be interpreted qualitatively since the actual results will differ considerably, depending on filter design, type of indicator, etc. It would appear from these curves that for best resolution with two different scanning speeds the filter band width should be designed for approximately the geometric mean between the two speeds. If signal-to-noise ratio should be of paramount importance the best value would be somewhat less than the geometric mean. It appears from available data that with good technique the resolution probably will not be degraded more than 10 per cent for scanning velocities ranging between 0.5 and 2 times the optimum. The degradation in signal-to-noise ratio for a corresponding range should be only a few decibels.

12.4.7 Location of Scanning Filter

Determination of the preferred frequency location for obtaining the desired characteristics in the scanning filter depends upon available technique in different parts of the frequency range and it is not possible to lay down any definite rules. Generally the location must be a compromise between the desire to use a relatively low frequency position, where a narrow band can easily be secured, and a higher location which would avoid one or more additional steps of modulation to bring the signals down to the filter frequency.

A location which would be advantageous purely from the standpoint of filter design is at zero frequency, since this would mean that the filter would need only one cutoff instead of two. However, if the signals were brought to zero frequency by rectification, it would be necessary first to separate them with a high-frequency filter affording a resolution substantially the same as the low-pass filter, so that no gain would result. It is possible that signals might be homodyned to zero frequency, but this has not been studied.

12.4.8 Desirable Scanning Filter Characteristics

The resolution[3] obtainable with scanning filters of various types was determined over a range of scanning velocities. The characteristics of the filter affording the best resolution may be summarized as follows:

1. The filter band width as measured between its 6-db discrimination points should be equal to approximately 1.2 times the square root of the scanning velocity. That is, $B = 1.2\sqrt{nF}$ where B = band width in cycles per second, n = the number of scans per second and F = the band scanned in cycles per second.
2. The attenuation outside the pass band of the filter should become asymptotic to about 30 to 40 db per octave of side-band frequency.
3. The response of the filter to suddenly applied pulses of a fundamental frequency equal to the mid-frequency of the filter should overshoot the steady-state value by 10 to 20 per cent.
4. The delay distortion across the pass band should be as small as is consistent with the above requirements.

The relation between steady-state characteristics and the characteristic given in (3) above is complex, depending on filter configuration. For example, in a filter with selectivity of 4 or 5 times 6 db per octave of side-band frequency, a square wave response exhibiting approximately 20 per cent overshoot is attained when the steady-state phase slope has a minimum at mid-band and maxima near the band edges, with a maximum-minimum ratio of 1.6.

12.5 PANORAMIC INDICATORS

The type of portrayal with which most of the discussion in this section will be concerned is that of signal presence versus frequency. Another type of display is one of signal patterns versus frequency, that is to say, an arrangement side by side of a number of facsimile patterns for different signals so as to permit simultaneous viewing. Other characteristics which may be shown are signal amplitude versus frequency, azimuth of signal source, etc.

By far the most desirable and most versatile form of indicator for a panoramic receiver is a cathode-ray tube. Further discussion will therefore be predicated upon this device except for specific treatment of other types of indicators.

12.5.1 Diagrams of Signal Presence versus Frequency

The simplest way of diagramming signal presence versus frequency on a cathode-ray tube is to cause the cathode-ray beam to swing in fairly rapid succession over a path of chosen shape, distance along which indicates frequency, with signals indicated either (1) by deflecting the beam sidewise with respect to the frequency trace or (2) by modulating the intensity of the beam.

Beam deflection affords a higher degree of resolution, the amount of the difference depending principally upon the design of the scanning filter and the level difference between input signals. On the other hand, beam deflection gives patterns which dance up and down because of signal modulation, static, etc., resulting in confusion and fatigue to the observer. Also if the frequency trace is folded on the tube face in any manner to obtain a longer frequency scale, beam deflection affords an opportunity for confusion between adjacent traces.

Beam modulation results in a diagram which is pleasing to the observer and gives less confusion between adjacent parts of a doubled-up frequency scale. The relative advantages of pleasing diagram as compared with superior resolution will depend on the circumstances involved. A disadvantage of beam modulation is that during periods when severe static crashes occur at frequent intervals the entire frequency scale is illuminated by each crash and observance of signals during periods between static crashes becomes quite difficult, especially if a persistent screen is employed.

Single-Line Trace

The simplest type of trace on which to show signal presence versus frequency is a single line produced by employing, for horizontal (or vertical) deflection, a wave corresponding to the frequency sweep. Such a single-line diagram is extremely satisfactory for many purposes. Sometimes, however, especially when scanning a wide band containing a large number of signals, the length of frequency scale obtained with a single line may be inadequate. In such cases a longer scale can be obtained, at a sacrifice of simplicity, in any of a number of ways. With any such long-scale arrangement, either beam modulation or beam deflection may be used to indicate signals, but in some instances beam deflection involves greater complication.

When stations occur at close intervals it is readily possible with available technique to display an amount of information far greater than can be appreciated. Hence the practical length of scale is usually fixed by observer capability. This varies considerably as between observers and depends upon the purpose for which the observations are made. Though no exact rule can be laid down, it appears that a single observer should not be required to monitor more than 15 or 20 stations at one time. If, however, it is permissible for the observer to shift his attention at intervals from one group of stations to another, a considerably larger number might be displayed.

Circular Trace

One way to lengthen the frequency scale is to employ a circular trace. This can be produced with an ordinary cathode-ray tube by applying to the vertical and horizontal deflecting plates or coils, respectively, two quadrature components of a sine wave whose frequency is the same as that of the frequency sweep. The arrangement is shown schematically in Figure 7. If it is desired to show signal presence by deflecting

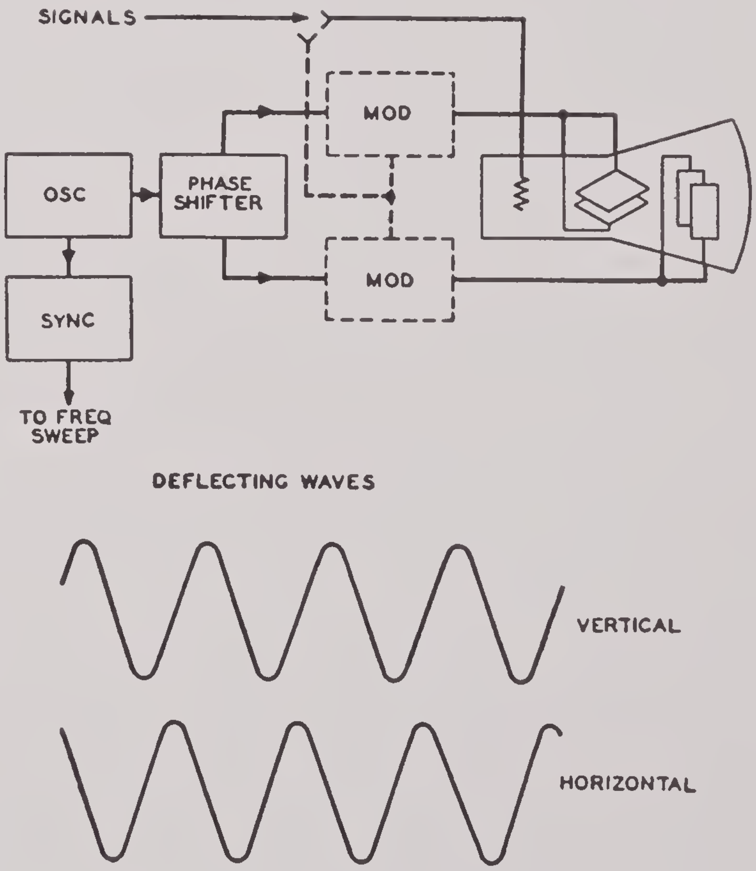

Figure 7. Method of producing circular frequency diagram.

the beam, this can be done by modulating the two quadrature components in accordance with rectified signal amplitude as indicated by the dotted lines. It is important in this case that well-balanced modulators with a minimum of unwanted intermodulation products be employed. Deflection to indicate signals may be toward or away from the center. An alternative way of producing the circular trace is to use mechanical rotation of magnetic deflecting coils.

SPIRAL TRACE

To produce a spiral trace,[1] the quadrature components which, if unmodulated, would give a circle are modulated by a sawtooth wave corresponding to the frequency sweep as shown in Figure 8. The number of turns in the spiral is determined by the ratio between the frequency of the sine wave oscillator and that of the sawtooth wave. An alternative method of producing a spiral is to employ a physically rotating magnetic field, the intensity of which is varied in accordance with the sawtooth frequency sweep.

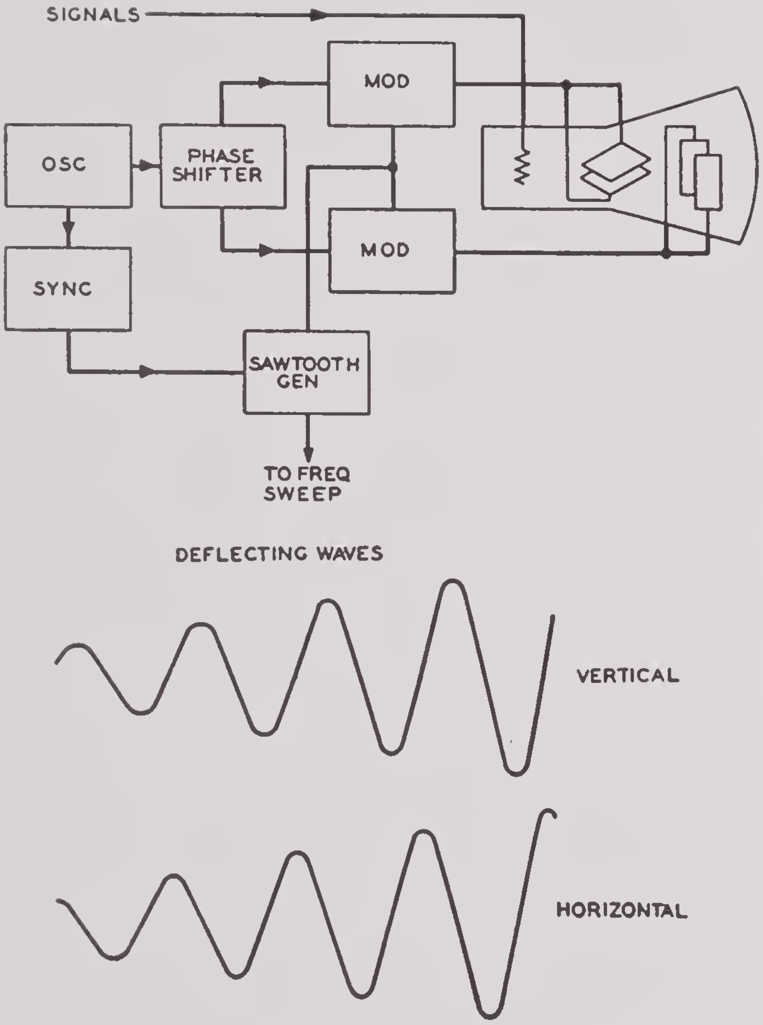

FIGURE 8. Method of producing spiral frequency diagram.

Modulation of the quadrature currents by the sweep sawtooth wave involves some difficulty if a reasonable approach to a true spiral is wanted. The sawtooth wave may be considered as composed of a number of sine wave components corresponding to the sawtooth frequency and its harmonics. In the modulation process it is necessary to preserve the side frequencies representing sawtooth components up to something like the tenth harmonic. Moreover, the nonlinear device which is used for the modulating process also generates harmonics of the quadrature sine wave frequencies. These harmonics lie in the same frequency band as the modulation products necessary for the transmission of the sawtooth wave and hence cannot be filtered out. Accordingly, careful design and balance of the modulating circuits are required. Balanced vacuum-tube modulators have been found satisfactory. The beam may be deflected to show signals by adding the rectified signals to the sawtooth modulating wave.

PARALLEL TRACES

A parallel-line diagram may be obtained by the method shown in Figure 9. In this case a sawtooth wave corresponding to the frequency sweep is applied to the vertical deflecting plates, and a sawtooth wave having a frequency equal to that of the vertical wave multiplied by the number of lines is used for horizontal deflection. The lines obtained with this arrange-

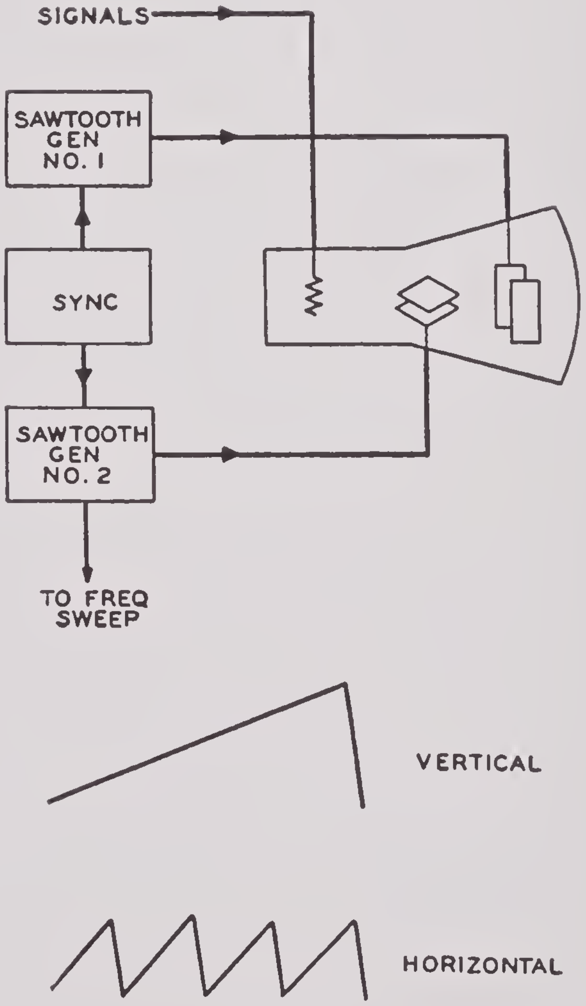

FIGURE 9. Method of producing parallel-line diagram.

ment, while parallel, are not exactly horizontal. To obtain horizontality the vertical deflecting wave must be a stepped wave as shown in Figure 10. To indicate signals by beam deflection the rectified signal wave is added to the vertical deflecting wave.

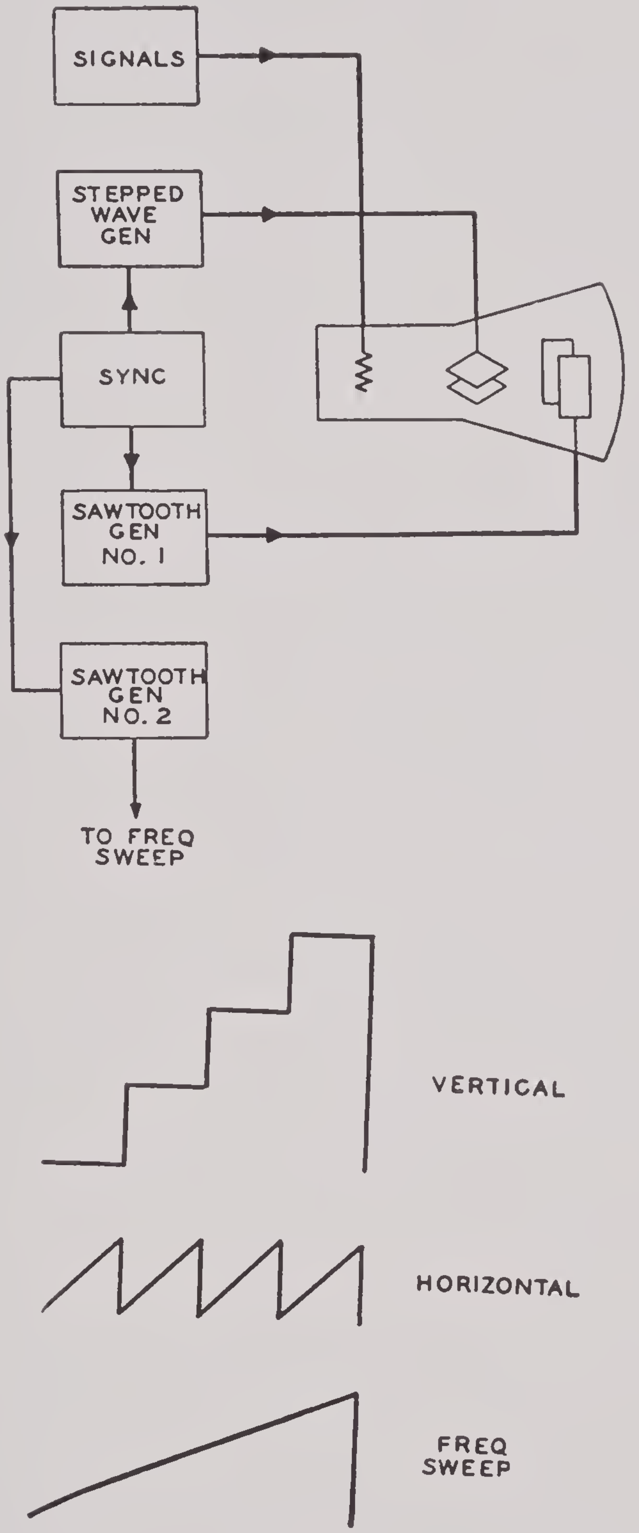

FIGURE 10. Method of producing horizontal parallel lines.

With a parallel-line diagram, the finite retrace time of the horizontal sweep means that signals may appear between parallel lines. If the retrace time is made sufficiently small, the retrace can be blanked out without entire loss of a signal, since any signal would appear in part at one or both of the opposite ends of adjacent lines. Theoretically it would be possible, by stopping or backing up the scanning oscillator during the retrace time, to avoid any loss whatever of frequency range during the retrace interval. This adds complications, however, and might introduce some degradation in receiver performance near the retrace.[b] See illustration No. ES-808183 in the C-36 final report, dated January 22, 1943, for a method of "backing up."

Retrace discontinuities during the scanning interval may be avoided by using a continuous zigzag scale, which may be produced by the method shown in Figure 11. The sawtooth wave which controls the frequency sweep is used for vertical deflection, while an isosceles triangular wave serves for the horizontal deflection. An objection to a zigzag form of diagram is

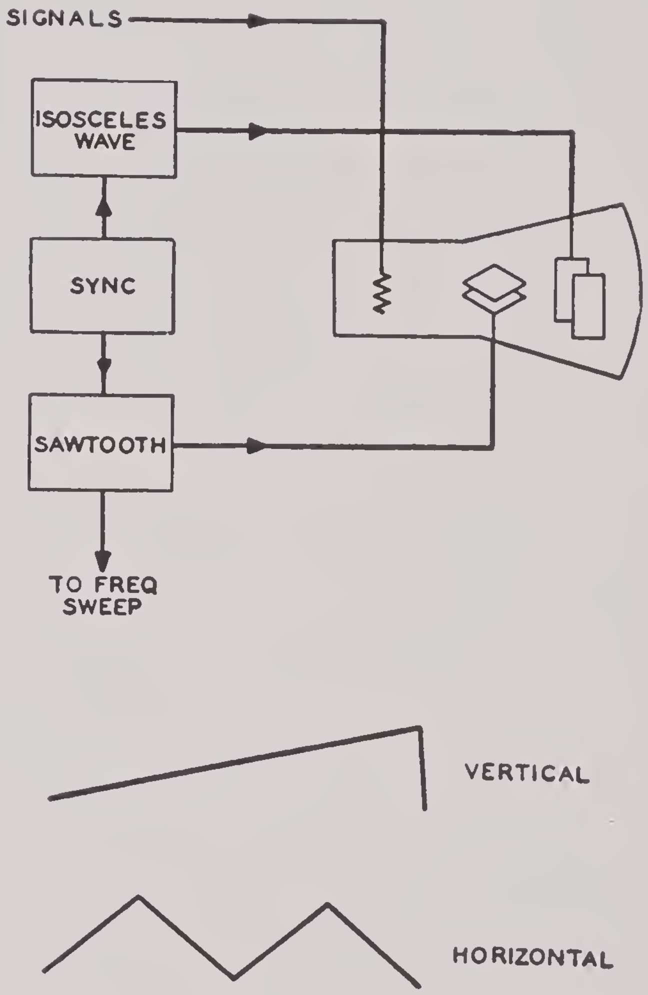

FIGURE 11. Method of producing zigzag diagram.

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[b] Cathode-ray tube indicators[1] having from 2 to 8 parallel-line traces on both low- and high-persistence screens were set up for rapid comparison with the spiral-trace indicator. Demonstrations were given for various members of NDRC, Army, and Navy. Personal preference varied, some preferring the spiral due to the absence of retraces in the scale, others preferring the parallel lines as being more nearly like printed page representation. The general conclusion seemed to be that for most applications there was little to choose between the types of traces. The high-persistence screen was preferred. It seems likely that each type of representation and various degrees of persistence may have advantages for certain special uses of panoramic receivers.

that since the spacing between lines is not uniform, there would be difficulty in distinguishing signals near the apexes of the zigzag.

12.5.2 Repetition Rates

The slower the repetition rate used for a frequency diagram, the narrower can be the scanning filter and hence the better the resolution. Too slow a rate, however, means less effective signal monitoring.

In earlier panoramic work flicker was considered to be an important factor in repetition rate and for this reason rates upwards of 15 cycles per second were employed. However, if a persistent phosphor of the cascade type is employed in conjunction with an orange viewing filter it becomes possible to employ repetition rates of less than about 4 cycles per second. Rates between 4 and 15 cycles per second are still somewhat objectionable as regards flicker.

The speed with which the presence of a signal can be detected is limited by the repetition rate. Requirements in this respect will depend upon the type and duration of the signals and upon the use being made of the panoramic receiver.

For keyed telegraph signals of the ordinary variety, the time is about equally divided between marking and spacing. In view of the short time required to pass a given frequency location the probability of encountering a marking signal in any one scan is only about 50 per cent. The probability of detecting a telegraph signal which is present in the frequency band accordingly depends upon the interval of observation and the repetition rate, as indicated by the following approximate formula:

$$P = 100(1 - 0.5^{n/T}) \, , \qquad (4)$$

where P = per cent probability of detecting telegraph signals during the observing period T (with T assumed to be $\geqq$ the period of one repetition) and n = repetition rate. Thus for $T = 1$ sec and $n = 1$ cycle per second, P equals 50 per cent. From this standpoint a repetition rate of less than 1 cycle per second may be undesirable.

Another disadvantage of an extremely low repetition rate is that even with a very long-persistence screen the complete signal diagram does not appear on the screen at one time. For certain types of work this might be objectionable. The minimum rate for which a complete diagram can be obtained with available phosphors is of the order of 1 cycle per second.

From all of the above it appears that the preferred range of repetition rates for frequency diagrams is from approximately 1 to 4 cycles per second, although far different rates may be used for special purposes if desired.

12.5.3 Type of Cathode-Ray Screen

Tests have indicated that a long-persistence screen is generally advantageous for frequency diagrams. Long persistence helps in keeping a signal of short duration on the screen long enough to permit observation and frequency determination. It greatly reduces flicker effects and eye fatigue and hence permits lower scanning rates with their inherently greater resolution and improvement in signal-to-noise ratio. Some further improvement in signal-to-noise ratio is obtained with a long-persistence screen as a result of cumulation of signal indications occurring at the same place on the screen, whereas noise traces occur at random. A disadvantage of a long-persistence screen, particularly when signal presence is indicated by spot modulation, is spreading of the phosphorescent trace. When the entire background is frequently illuminated by static crashes, long persistence is disadvantageous since it tends to prevent observation of signals by means of the fluorescent patterns between crashes.

In practice a cascade screen employing a P7 phosphor with blue-white fluorescence and orange phosphorescence has been found quite satisfactory. By providing both orange and blue viewing filters either the fluorescence or the phosphorescence, as desired, may be masked to a large extent.

12.5.4 Indicator Resolution

The resolution obtainable in a frequency diagram can be no better than the fraction of the total scanning band which corresponds to the ratio of the resolving power of the indicator to the length of frequency scale. The resolving power of a cathode-ray tube indicator is determined by the diameter of the spot. While extremely small spots can be realized under laboratory conditions, practical operating results are generally poorer. With a long-persistence screen the spot tends to spread, thus reducing resolving power. Size of spot increases less rapidly than tube size, so that there is an advantage in going to larger tubes. When intensity modulation of the spot is employed, the resolution is much poorer because of defocusing or blooming of the spot as the beam current becomes large. When beam-intensity modulation is used the size of spot depends on the maximum beam intensity. As the "blooming" point is approached, the size of spot increases very rapidly.

If the frequency scale is long enough so that the resolving power of the indicator is not a direct limitation, the capability of the indicator for distinguishing between two nearby signals still affects the scanning resolution. With the beam-deflection type of indicator, the signal takes the form of an inverted V or U according to the filter characteristic. Adjacent signals are distinguished by observing a slight dip between two inverted V's or U's which are not quite superimposed. For certain repetition rates, particularly in the range 5 to 15 cycles per second, a more sensitive discrimination index can be found in a beat which occurs between the two signals at the repetition rate. Under practical conditions, however, particularly when keyed signals and noise are present, this beat method of discrimination is of little value.

With intensity modulation of the cathode-ray beam, signals are distinguished by differences between the size and shape of adjacent spots. Discrimination in this case is considerably poorer than that with beam deflection.

12.5.5 Facsimile-Type Diagrams

An alternative to a diagram of signal presence versus frequency is one in which facsimile-type patterns for a number of signals of different frequencies are arranged side by side for simultaneous viewing. One way of producing such a diagram,[2] is to employ a special type of cathode-ray tube which is rotated so as to give a continuous motion of the fluorescent screen in relation to the scanning beam. The sawtooth wave which controls the frequency sweep is employed to deflect the cathode-ray beam in one dimension across the screen, the moving screen makes time the other dimension, and signal amplitude modulates the beam intensity. Signal patterns are seen as a result of screen persistence.

Low brightness of aftertrace is a limitation in the moving-screen tubes thus far constructed. This makes it necessary to observe in an almost completely darkened room. However, considerable improvement in this respect is believed possible.

With a continuously rotating cathode-ray screen, there is a possibility that the phosphorescent traces after completing one revolution will not have decayed sufficiently to avoid interference with the new traces. This might be avoided by accelerating the decay of the stored energy subsequent to its passage across the viewing window. For this purpose infrared rays could be used, which would free the stored energy largely as heat but partly as accelerated light emission. Although it appears possible to develop a satisfactory arrangement for wiping out the phosphorescent trace in this way, this has not been found necessary with the particular type of moving screen tube thus far employed.

An alternative method of obtaining a panoramic diagram of the facsimile type is to substitute for the moving cathode-ray tube a moving web of phosphorescent material activated by a light beam whose position is adjusted in accordance with the frequency scan and whose intensity is modulated by the input signals.[c]

12.5.6 Recording Radio Telegraph Signals

With a suitably designed moving-screen indicator observation of frequency patterns is possible. Duration of transmission can be observed, which may permit the pairing up of transmissions of two stations. Telegraph code with hand sending can be reproduced and read from the screen. By using a very narrow frequency sweep the varying frequency limits and syllabic intensity changes of a-m speech side bands can be observed.

For observation of such signal patterns there is an optimum rate of travel of the moving screen, since too slow a rate does not afford adequate resolution and too rapid a rate makes it impossible for the eye to follow the pattern. Satisfactory results have been obtained with a rate of screen travel of about 1 in. per second. To resolve telegraph code it is necessary that the repetition rate of the frequency scanning be sufficiently high to provide several scannings during the interval of a single telegraph dot. A repetition rate of 60 cycles per second has been found satisfactory for hand telegraph speeds. Repetition rates much higher than this would afford little advantage for continuous reading of high-speed telegraph in view of the limitations of vision.

The keying speed[1] of the fastest transmission that is to be recorded determines the minimum scanning frequency for the receiver. If it is assumed that there should be 3 scans per dot, the number of scans per second (n) should be six times the keying fundamental frequency of 2.4 times the number of words per minute.

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[c] In November 1945, the Bell Telephone Laboratories demonstrated such a moving-web system designed to aid deaf persons to learn to talk by viewing moving visible images corresponding to speech sounds.

The primary relationship underlying scanning receiver operation may be represented by

$$S = K_S\sqrt{nF},$$

where S = signal resolution or, in this case, minimum frequency difference between adjacent stations in cycles per second.

K_S = a constant.

n = number of scans per second.

F = band width scanned in cycles per second.

This means that the band width that may be scanned varies inversely with the number of words per minute of the highest speed signal to be recorded. Also the minimum allowable frequency separation between signals increases as the square root of either the number of words per minute or the band width scanned.

In conditions where the stations are uniformly spaced, we can rewrite the above equation as $S = K_S^2nx$, where x equals the number of stations recorded. In other words, the number of equally spaced stations that may be simultaneously recorded varies directly with their frequency separation and inversely with the number of words per minute transmitted by the fastest station.

As the relative amplitude increases, K_S increases in a complicated manner, for example, relative amplitudes of 30 db may mean that K_S is three times what it would be for equal signals, i.e., the number of equally spaced signals that could be recorded would be reduced ninefold.

Another element controlling the value of K_S is the type of indicator used in the receiver. The indicator resulting in the lowest values of K_S at the present time is an amplitude deflection indicator which is not suitable for recording. The present best indicator for recording (the so-called spot or intensity indicator) causes several-fold sacrifice in K_S. K_S is then contingent on the amplitude ratio between the weakest and the strongest signals in the band scanned, rather than merely on the ratio between adjacent stations.

Table 1 illustrates the interrelationship of the factors discussed above. K_S for equal signals is assumed to be 1.25 and K_S for signals of 30 db relative strength is assumed to be 3.6, although these values are somewhat smaller than could probably be realized under service conditions.

12.5.7 Other Types of Diagrams

For some applications, as for example, in identifying the character of signals or in matching signal strengths for different purposes, a diagram of amplitude versus frequency of radio signals is useful. Such a diagram is afforded by a cathode-ray tube using beam deflection as already described, provided limiting is not introduced ahead of the indicator. It should be noted that the resolution obtainable in a diagram to be used for amplitude versus frequency determination is many times poorer than that obtainable when the requirement is merely to indicate what frequencies are present. In the latter case the transient responses before and after the main response from a signal may blend with the response from another signal to alter greatly its apparent magnitude without preventing

TABLE 1. Effect of variables on scanning-receiver recording performance.

Minimum station separation (kc)	Maximum words per minute per station	No. of scans per sec (n)	Max. band scanned in kc for relative signal strength indicated		Number of equally spaced signals (x) for relative strength indicated	
			0 db	30 db	0 db	30 db
5	30	72	222	27	44	5
	50	120	133	16	27	3
	100	240	67	8	13	2
	200	480	33	4	7	1
	250	600	27	3	5	1
10	30	72	890	107	89	11
	50	120	534	64	53	6
	100	240	267	32	27	3
	200	480	133	16	13	2
	250	600	107	13	11	1
20	30	72	3,560	427	178	21
	50	120	2,130	256	107	13
	100	240	1,070	128	53	6
	200	480	535	64	27	3
	250	600	427	51	21	3

observation of its presence. When amplitudes are to be depicted accurately the signal separation must be sufficiently large to prevent intermingling of transient responses.

The azimuth of a number of radio signals of different frequency can be displayed simultaneously by means of a panoramic receiver. Such a diagram might be in the form of a straight line, distance along which represents azimuth, but more conveniently would take the form of a circle on which azimuth is represented by angular position and signal presence is indicated either by beam modulation or beam deflection toward or away from the center. A method of obtaining such a diagram, using beam deflection, is shown schematically in Figure 12. In addition to the frequency scan it is necessary that the receiving antenna scan azimuth in any of various well-known ways. One scanning rate must be considerably higher than the other for adequate resolution. A similar arrangement might be used to provide a diagram of azimuth of signal source versus frequency. This might be a polar diagram in which frequency is represented by radial distance and azimuth by angular position, using the schematic arrangement of Figure 13.

12.5.8 Mechanical Forms of Indicators

Mechanical counterparts can be devised for most of the types of diagrams which have been described. In general these are obtained by (a) providing a source of light which is modulated in accordance with the signal intensities delivered by the scanning filter and (b) viewing this source of light through an aperture the position of which is mechanically varied synchronously with the frequency scanning. The source

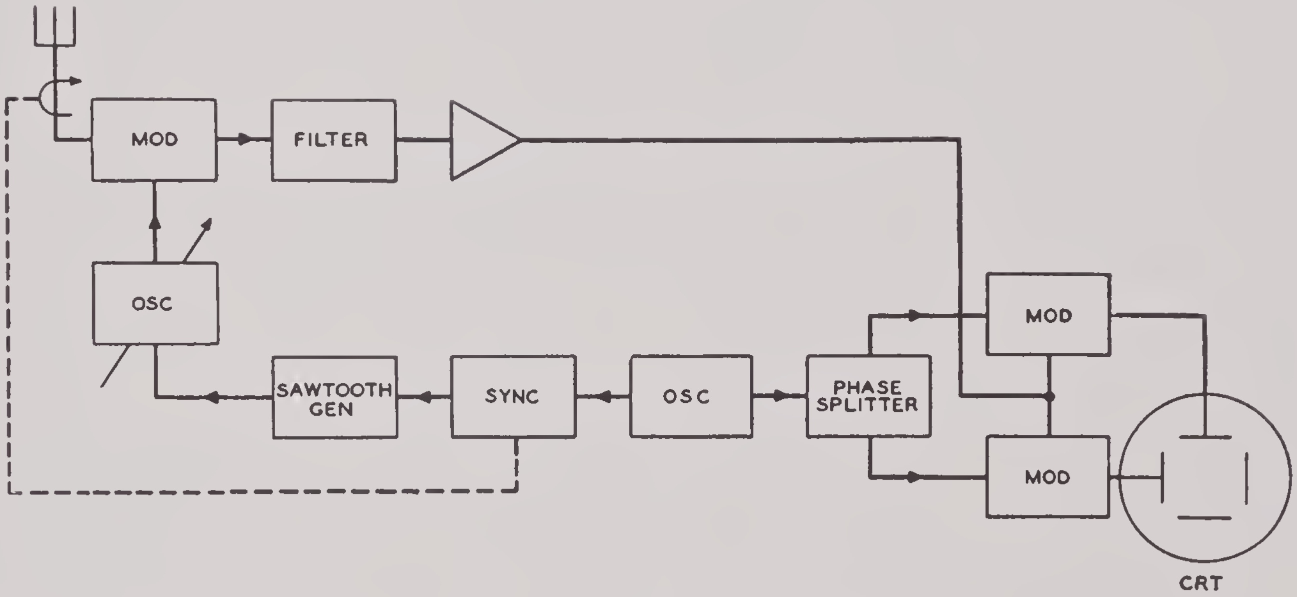

Figure 12. Method of producing diagram in which azimuth is represented by angular position on circle.

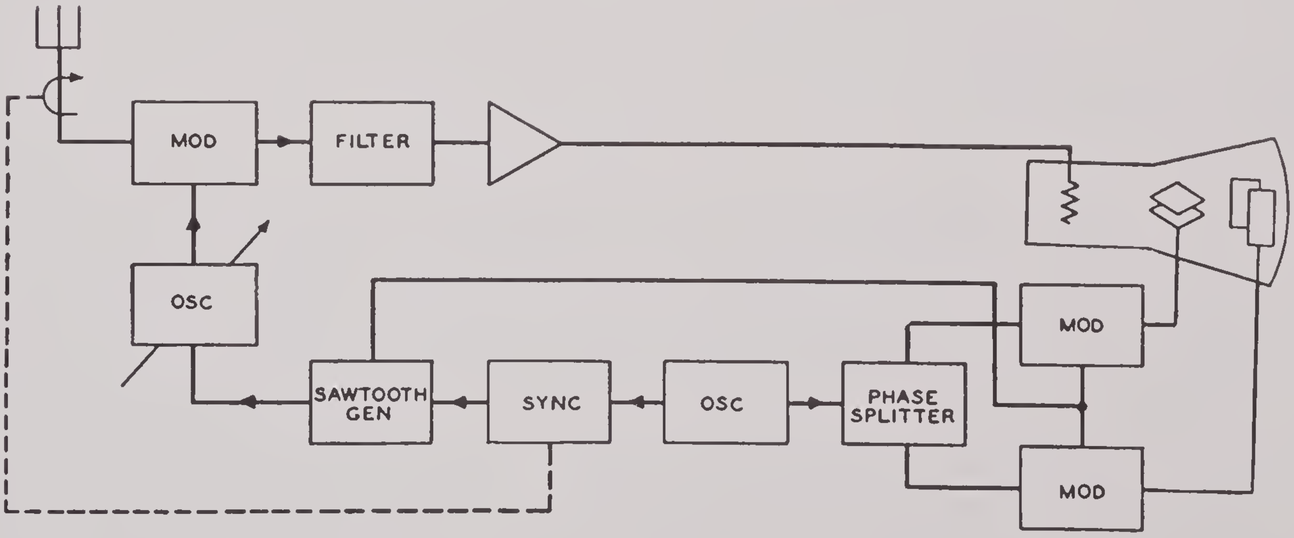

Figure 13. Method of producing circular diagram in which frequency is shown by distance along radius, and azimuth by angular position.

of light may be, for example, a neon tube. If this source is arranged so as to illuminate a line which is viewed through a rotating cylinder provided with a helical slit, a single-line diagram of signal presence versus frequency can be obtained. Similarly by illuminating a circular path around which the viewing aperture moves, a circular frequency diagram can be obtained. By illuminating an area and using suitable viewing arrangements, spiral or other types of diagrams can be obtained.

Mechanical devices of this kind may be smaller than cathode-ray indicators but seem likely to involve considerable sacrifice in resolution.

12.5.9 Alarms—Automatic Stopping

When the number of signals in the scanning range is not large, a convenient addition to a scanning receiver may be an audible alarm which sounds either momentarily or continuously whenever a signal is encountered. In some instances it may be useful to provide means for automatically stopping the scan upon encountering a signal, in order to permit examination of it, or to indicate its presence or other characteristics if it disappears quickly.

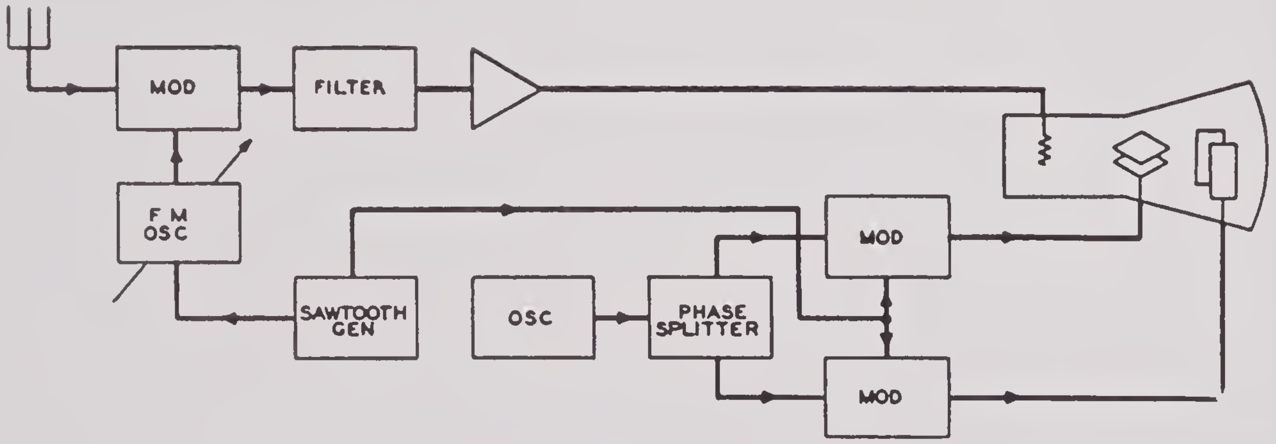

Figure 14. Method of producing circular signal traces.

12.6 SENSITIVITY

The signal-to-noise ratio obtainable in a well-designed panoramic receiver is necessarily poorer than that obtainable with an ordinary aural receiver. This is because the width of the selecting filter as determined by scanning limitations is greater than that needed for fixed tuning. In the case of telegraph the difference in signal-to-noise ratio may range from 10 to 30 db, depending on the scanning speed. For a-m speech transmission the difference is smaller and for f-m transmission it may be practically negligible.

If the scanning speed actually used differs from the optimum for which the scanning filter was designed, there will be a sacrifice in signal-to-noise ratio as discussed above.

It is believed that the capabilities of the eye in picking out signals in the presence of noise are somewhat poorer than those of the ear. Signal-to-noise discrimination possible in visual reception varies considerably, depending upon the type of indicator and the type of diagram. If a persistent phosphor is used, cumulation may be obtained for successive signals while noise is distributed at random. From this standpoint beam modulation is somewhat superior to beam deflection.

It is possible that a slight gain in signal-to-noise discrimination can be obtained with a diagram in which each signal is made to appear as a line trace, either by moving the screen with respect to the cathode-ray beam or vice versa. Thus a time pattern can be obtained with a fixed cathode-ray tube by moving the beam so that the trace corresponding to any signal is a line which may be circular, vertical, or horizontal. A method of producing circular signal traces is illustrated in Figure 14. This employs a radial frequency sweep and circular time sweep, the radial scanning rate being a high multiple of the circular sweep. A method for producing vertical line traces is shown in Figure 15. In this case the horizontal sweep corresponds to the frequency scan while the vertical sweep is much slower. With any of these schemes for producing line traces, it appears likely that there may be an optimum trace velocity which yields the best discrimination between signal and noise.

12.7 FREQUENCY DETERMINATION

Generally it is desirable in a panoramic receiver to be able to determine the frequency of observed signals with considerable accuracy. This can be done in a number of ways.

Fixed frequency markings can be provided on the face of the tube and arrangements in the receiver to align the frequency scan with these markings. Usually it is not possible to obtain sufficient accuracy with this method.

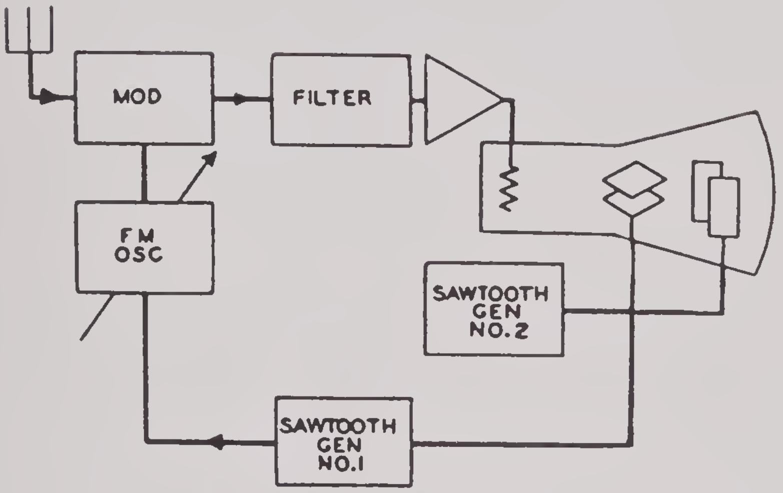

Figure 15. Method of producing vertical line traces.

A complete frequency scale may be produced by electrical markings on the screen. Even though the electrically produced scale markings are precise, it may still be difficult to obtain the requisite accuracy in reading the frequency of observed signals by interpolation between markings.

The frequency-marking method which has been found of greatest utility is to provide a signal generator whose frequency may be adjusted over the scanning band. This makes it possible to produce on the indicator a marking signal whose frequency can be made to coincide with that of any received signal. Such a signal generator may be advantageously combined with an aural receiver arranged to monitor any signal received by the panoramic receiver. In this case the marking frequency is made the same as the frequency to which the aural receiver is tuned. This permits tuning the aural receiver quickly to any desired signal trace on the screen and reading the frequency of that signal on the receiver dial.

To avoid confusion the marking frequency may be indicated in such a way as to differentiate it from the received signals, for example, by using upward deflection for signals and downward deflection for the marker. However, the use of different directions of deflection for signals and marker results in somewhat poorer scanning efficiency. Some of the scans (perhaps 25 per cent) must be devoted to scanning the marker alone. This means that the band scanned must be reduced (perhaps 25 per cent) if the same resolution is to be obtained.[d]

With a receiver of the heterodyne type, the frequency of the beating oscillator differs from the input frequency by the intermediate frequency. Accordingly

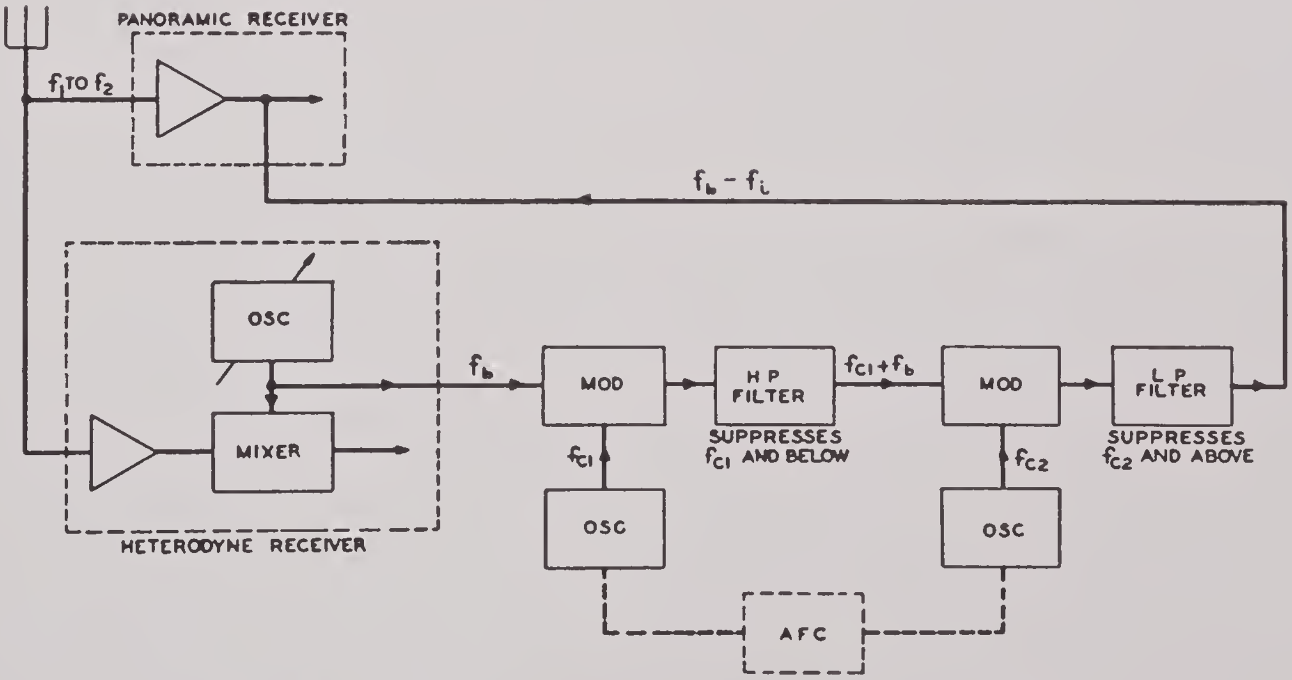

Figure 16. Derivation of marker by double-modulation method.

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[d]Other tests[1] were made in which the marker, a manually controlled signal used to identify an unknown signal, was scanned alternately with the signals received from the antenna. This permitted the marker to be adjusted much closer to the exact frequency of the unidentified signal but was not an appreciable aid in tuning in the latter on the monitoring receiver. In many applications this arrangement has the objection that it results in either fewer indications per second, narrower band scanned, or poorer resolution. In applications to frequency matching for "jammers" the additional accuracy far outweighs the objections.

a frequency corresponding to the receiver tuning may be produced from the beating-oscillator frequency. There are two general methods:

1. A separate oscillator ganged to the beating oscillator and differing in frequency by the intermediate frequency. The difficulty in this case is to obtain accurate tracking over the entire frequency range.
2. Electric frequency derivation. If this is done by combining the beating oscillator and the intermediate frequency in a mixer, unwanted frequencies are present close to the desired frequency. Methods of eliminating or avoiding such unwanted frequencies are as follows:
    a. Ganged tuning.
    b. Use of double modulation as illustrated in Figure 16.
    c. Use of automatic frequency control as illustrated, for example, in Figure 17.
    d. Some sort of single side-band modulation scheme using phase balance or a combination of amplitude and frequency modulation might be possible.

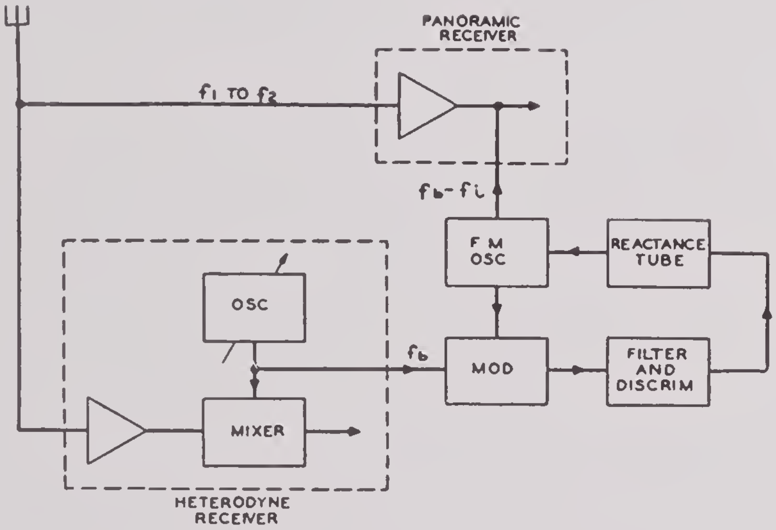

FIGURE 17. Derivation of marker by automatic-frequency-control method.

12.8 BLANKING OUT RECOGNIZED STATIONS

When the number of signals appearing on a panoramic indicator is large, observation can be facilitated if identified signals are blanked out in some manner so that the sudden appearance of a new signal can be readily detected. There are several ways of doing this.

Rejection or trap circuits, preferably adjustable, might be used ahead of the scanning oscillator to suppress recognized signals. If a number of different signals are to be suppressed, this scheme involves considerable complication. Also it would be difficult to obtain sufficiently sharp discrimination to avoid at least partial suppression of frequencies on either side of the desired signal. The scheme may be useful, however, for suppressing a small number of high-level stations.

Another way of blanking out recognized stations would be to generate a number of different blanking pulses, adjustable in time position and width, which could be applied directly to the cathode-ray tube. This likewise involves considerable complication.

A simple scheme which has been used with the intensity type of indicator is to apply black paint to the face of the tube to cover up identified signals. With a suitable paint the screen can be wiped clean at intervals as desired.

A scheme which has been suggested for blanking out is to record incoming signals on a continuous magnetic tape and utilize recorded signals to balance out incoming signals so that only a signal not previously recorded would appear on the screen. This scheme appears to involve certain practical difficulties. An alternative scheme which might have advantages would be to use a magnetic tape on which blanking signals are produced by a local signal generator.

Blanking schemes necessarily involve some sacrifice in resolution in the blanking region. A further disadvantage of any blanking-out scheme is that an identified signal may disappear and be replaced by an unidentified one without the observer being aware of this. Also, when blanking out is used a low-powered signal which is adjacent to a high-powered known signal may pass unobserved, whereas otherwise it might have been noted during idle periods of the larger signal.

12.9 NARROW-BAND SCANNING — RANGE EXPANSION — COMBINED SCANNING

A panoramic arrangement for scanning a relatively narrow band on either side of the tuning frequency has been found to be a useful attachment for an ordinary heterodyne receiver.[15] The arrangement is illustrated schematically in Figure 18. The input to the panoramic unit is derived from the i-f output of the mixer, which should be reasonably flat over the sweep range. The signals from the mixer circuit are swept past a scanning filter by means of a beating oscillator which is swung through a range of about ±50 kc. The diagram normally is in the form of signal amplitude versus frequency.

A narrow-band scanner of this kind may be combined with a wide-band panoramic receiver to obtain both a low-definition wide-band and a high-definition

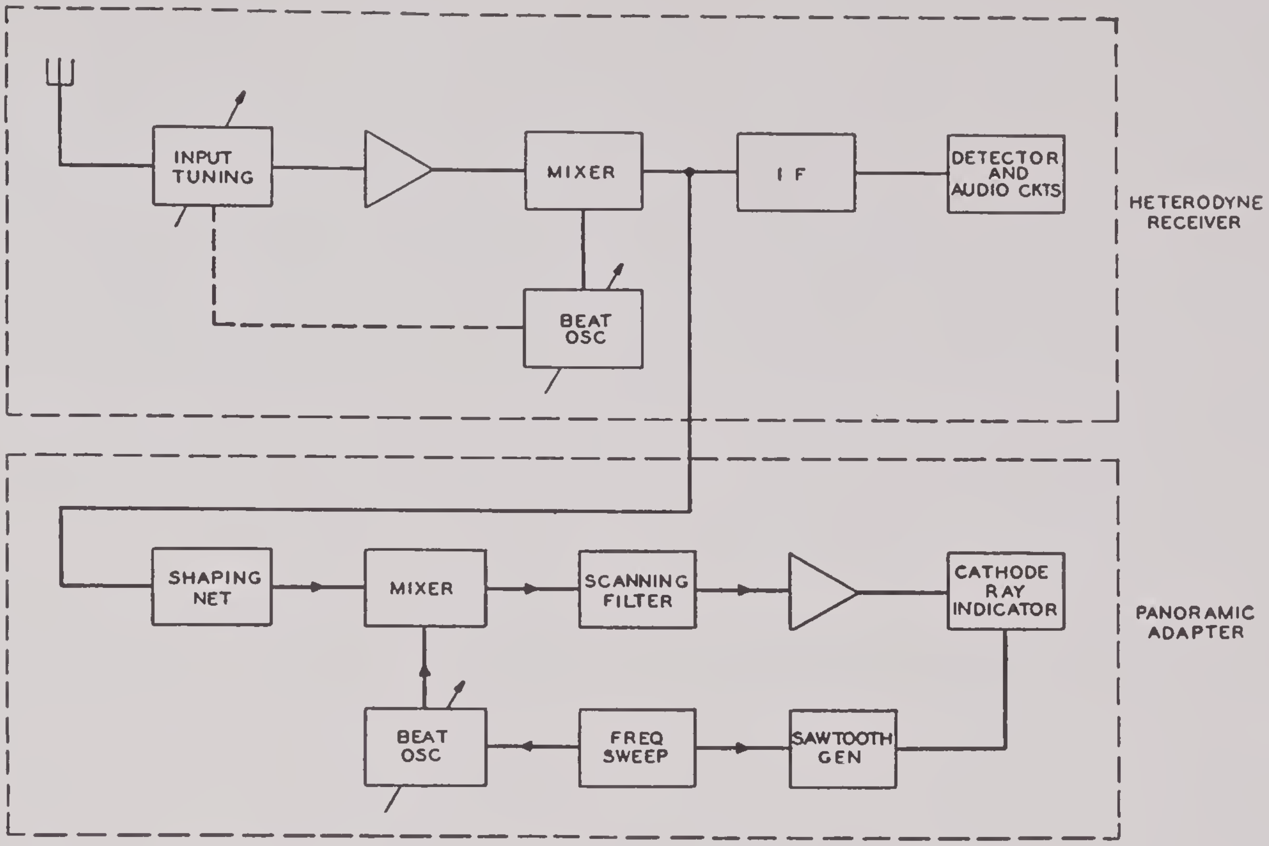

FIGURE 18. Schematic of panoramic adapter.

narrow-band diagram. The narrow band may be moved manually or automatically over the wide band.

A combination wide- and narrow-band diagram may be provided on a single cathode-ray tube, the narrow band representing merely an expansion of a certain part of the wide band. When the resolving power of the cathode-ray tube itself is the limiting factor, better resolution is obtained merely by expanding a part of the frequency scale, as shown, for example, in Figure 19 for a single-line frequency diagram. The wave used for horizontal deflection, instead of being merely a simple sawtooth, is a combination of a sawtooth and a stepped wave with a steep slope between steps. The position of this slope would be adjustable so that any part of the frequency range could be expanded.

The above scheme is of little value if the resolution is limited, not by the indicator, but by the scanning filter. In this case it is necessary to obtain greater scanning resolution in the narrow band. The best way to get this is to use separate wide- and narrow-band scans. Results may be displayed as two diagrams, the wide- and narrow-band pictures may be combined on one screen. With the arrangement of Figure 20, alternate wide- and narrow-band scans may be shown on separate lines of the cathode-ray tube. Instead of using two separate scans, it would theoretically be possible to have a single wide-band sweep with some arrangement for slowing down and introducing a narrower filter over a small part. The practical value of such a scheme is doubtful.

One way of improving the resolution of a panoramic receiver is to provide a number of narrow-band scanners positioned side by side. The equivalent of this can be had, without complete duplication of equipment, by providing a number of filters spaced apart over a wide band and sweeping the wide band through a frequency interval corresponding to the spacing between filters. The outputs of the several filters would then be scanned at a rate which is a large multiple of the rate of input frequency swing. The method is illustrated schematically in Figure 21. To produce a single-line diagram of signal presence versus frequency a combination wave consisting of a sawtooth corresponding to the input swing and a stepped wave corresponding to the filter-output scan would be used for horizontal deflection of the cathode-ray beam. Signals might be indicated by beam deflection or beam modulation. Assuming a given total band, the increase in resolution obtained by dividing up with filters according to this method increases as the square root of the number of filters. On the other hand, if a given band can be scanned with a certain resolution using one filter, this band increases directly with the number of filters.

12.10 RECORDING

Strictly speaking, a recording receiver is not a panoramic receiver and therefore lies outside the scope of this report. Application of recording to receivers of the panoramic type will, however, be briefly considered.

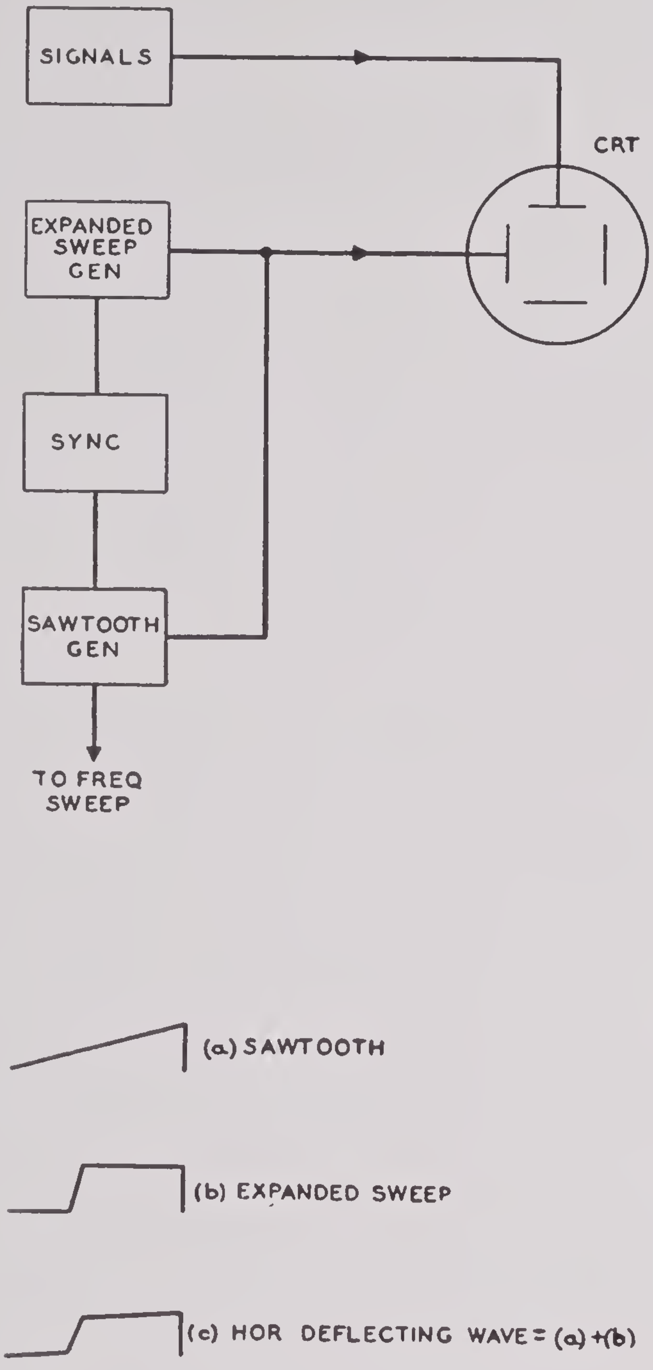

Figure 19. Method of expanding part of frequency scale.

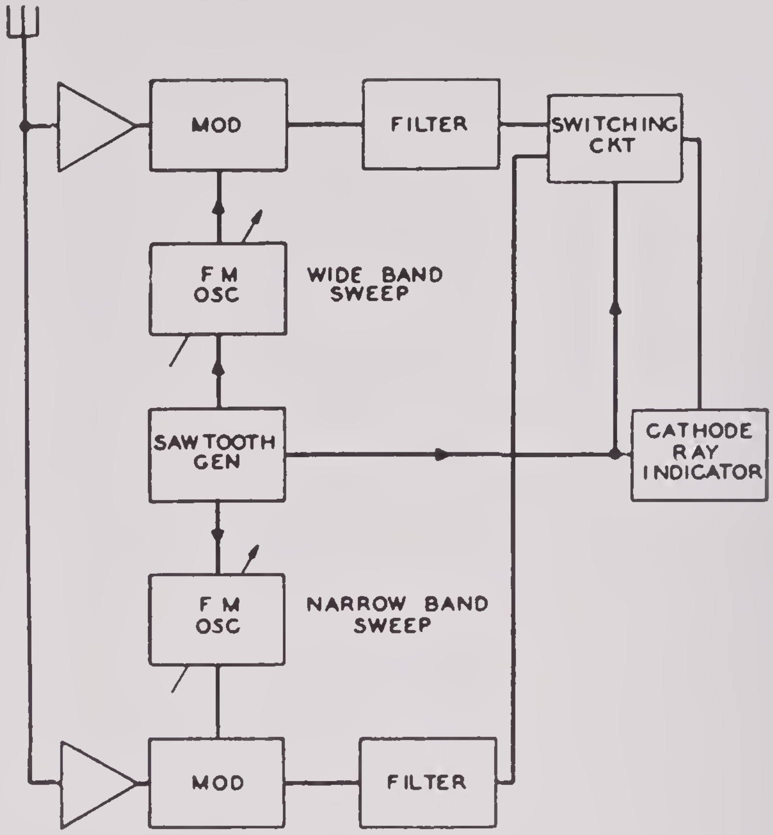

Figure 20. Combined wide- and narrow-band diagrams.

Principal interest probably resides in facsimile-type recording where a number of signal patterns for stations of different frequency are arranged side by side. For this it is necessary to have a moving record activated by a light (or electronic) source whose intensity is modulated to indicate signals and which is moved transversely across the record in synchronism with the scanning of the frequency range. A suitably high-speed facsimile recording method (electrolytic, electrothermal, photographic, etc.) may be employed. The receiving equipment proper may be the same as for a panoramic receiver.

Although a panoramic facsimile recorder avoids the limitations of the eye which have been noted for a facsimile type of indicating system, some important limitations remain. If, as is usually the case, it is desired to resolve telegraph code, it is necessary to employ a fairly high repetition rate for the scanning. The band width that may be scanned varies inversely with the number of words per minute transmitted in the highest speed signal to be recorded. Conversely, the minimum allowable frequency separation between signals to be recorded increases as the square root of either the number of words per minute of the highest speed signal or the band width to be scanned. A repetition rate of about 60 cycles is desirable for hand telegraph speeds and correspondingly higher rates for higher speeds. As already discussed, the use of a high repetition rate impairs frequency resolution. Furthermore, the speed of the moving record is determined by the resolution desired along the time axis, which again is dependent on telegraph speed. A speed of the order of 1 in. per second is probably suitable for hand telegraph speeds. One inch per second means 1⅓ miles of record per day, so that analysis of the record becomes difficult. If, in view of the limited frequency band that can be covered by a single recorder, a number of recorders are employed for adjacent frequency bands, the problem of record analysis is correspondingly multiplied.

It has been proposed that length of record be reduced by employing a combination of panoramic indicator and recorder, making a record only when signals of interest are observed on the indicator. This requires that the signals be stored long enough to permit a

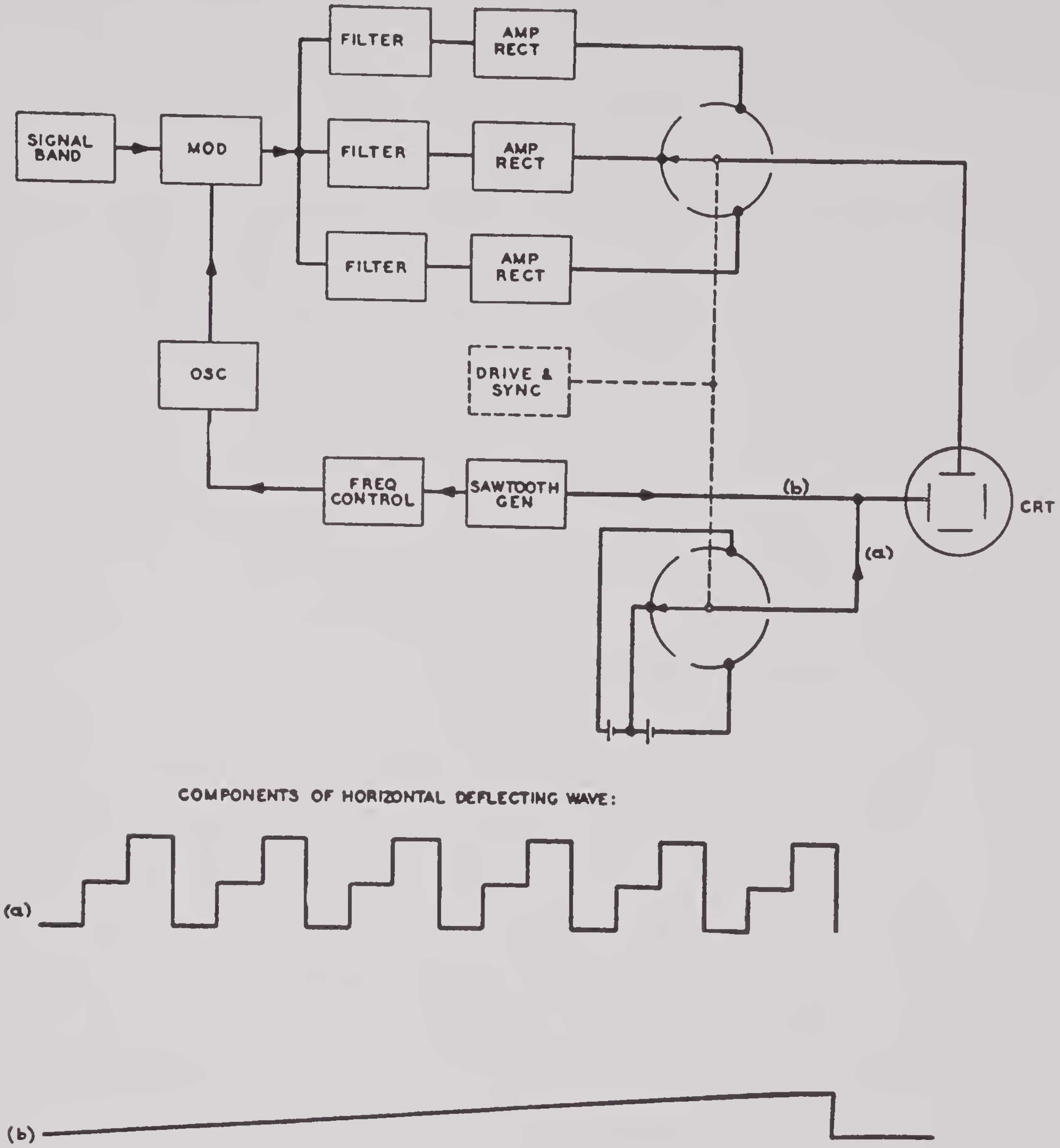

Figure 21. System with multiple scanning filters.

decision as to the desirability of recording. Such storage may be obtained by recording the signals on a magnetic tape.

12.11 PANORAMIC PULSE RECEPTION

Panoramic reception of radar system pulses differs in certain respects from reception of ordinary signals. In the first place, the frequency spectrum of a high-frequency pulse is quite broad.[16] For practical purposes most of the energy of the pulsed high frequency may be assumed to lie within a band centered at the carrier frequency and having a width equal to twice the reciprocal of the pulse length. Thus the band for a 1-μsec pulse would be approximately 2 mc.

It is desirable that the scanning filter be as wide as the pulse signal band as roughly defined above. This is for two reasons: First, this affords good signal-to-noise ratio; and second, it permits stopping the receiver and viewing pulse shape. Since the resolution is limited in any case by signal band width, no sacrifice of resolution results from making the filter band at least as wide as the signal band.

If the width of the scanning filter is made equal to the pulse signal band, then it turns out that for practical repetition rates and total frequency ranges the scanning speed is no longer a limitation on resolution. Thus, for example, with a 2-mc filter the scanning speed (product of repetition rate and total frequency band) may be of the order of 10^{12} cycles per second.

A further factor in scanning reception of pulse signals is the low duty cycle, i.e., the fact that the signals are present only a very small part of the time. In the processes of scanning reception a number of pulses are lost, the fraction received being equal to the ratio of the width of scanning filter to the total frequency band covered. If, in addition to frequency scanning, the receiver employs azimuth scanning or if the source of the pulse signals employs azimuth scanning, the fraction of the original pulses received is much further reduced.

Pulses must be present during a sufficient percentage of the time to actuate the panoramic indicator. The minimum band width for the scanning filter to satisfy this condition may be stated as

$$B = K_I \frac{F}{N}, \tag{5}$$

where B is the band width, F is the total frequency band, N is the pulsing rate, and K_I is the number of pulses that must be received per second to actuate the indicator. If, for example, the total band to be swept were assumed to be 400 mc, if the pulsing rate were 400 per second and if 10 pulses per second were necessary to actuate the indicator, then the scanning filter should be 10 mc wide. While a complete discussion of these factors is beyond the scope of this report, it is evident that if very wide frequency bands are to be scanned, it becomes necessary to make the scanning filter considerably wider than the band width of the pulse signal. Also it is clear that scanning reception in combination with azimuth scanning in the receiver or in the transmitter becomes extremely difficult.

The signal-to-noise ratio, and hence the realizable sensitivity, of a scanning receiver for radar pulses is poorer than that of the same receiver for continuous narrow-band signals of the same field strength. There are two reasons for this. First, the pulse receiver is responsive to noise all the time while the pulses are present only a small part of the time. Secondly, the received noise power varies directly with band width, so that the noise is correspondingly greater in receiving wide-band pulse power than in receiving the same power concentrated in a narrow band.

12.12 PANORAMIC RECEIVERS WITHOUT FREQUENCY SWEEP

Frequency sweep, which has formed the basis of a large part of the material in the foregoing sections, is not essential for panoramic reception. An alternative is to provide a selective device or devices whereby the input signal band is separated into small component bands which are applied in succession to a visual indicator. The selection may be accomplished by a series of filters having a common input and separate outputs. The filter bands may be made adjacent, contiguous, or slightly overlapping as desired. These outputs may be scanned in rapid succession and resultant signals used to provide the indicator diagram. A possible arrangement is shown schematically in Figure 22. A method of this kind has been used in studying the effect in a short-wave radio-telephone channel of changes in the transmission medium[17] and numerous subsequent applications have been made.[18] The same general method may be employed for facsimile or other kinds of recording.

The outputs of the multiple selecting circuits may be scanned at practically any desired rate if amplifiers or other circuits which follow this scanning are made wide enough. Consequently, by eliminating the frequency sweep the problems inherent in the scanning filter are avoided. To provide a sufficient number of filters to cover the desired frequency band with adequate resolution may, however, involve considerable complexity of equipment.

12.13 USES OF PANORAMIC RECEIVERS

Listed below are possible types of uses, both military and nonmilitary, for panoramic receivers. While considerable information is available regarding possibilities of panoramic receivers for some of these applications, much further study and experience will be necessary for a complete appraisal.

Intercept. Panoramic reception may serve as an adjunct to aural reception in different types of intercept work, including particularly intelligence and traffic studies. It may supplement, or possibly avoid, continuous aural searching. Wide- or narrow-band scanning may be used to provide information on signal presence, signal frequency, signal characteristics, duration of transmission, etc.

Communication. In communication channels or networks, panoramic reception may be used in adjusting the frequency of transmitters to assignments which are suitable from the standpoint of other signals and interference. Where transmissions are to be received on different frequencies over a wide band, a panoramic receiver could function as a call indicator. Narrow-band scanning would facilitate detection of, and receiver adjustment for, moderate variations in transmitter frequency. Panoramic reception might be used to detect interference or jamming by the enemy,

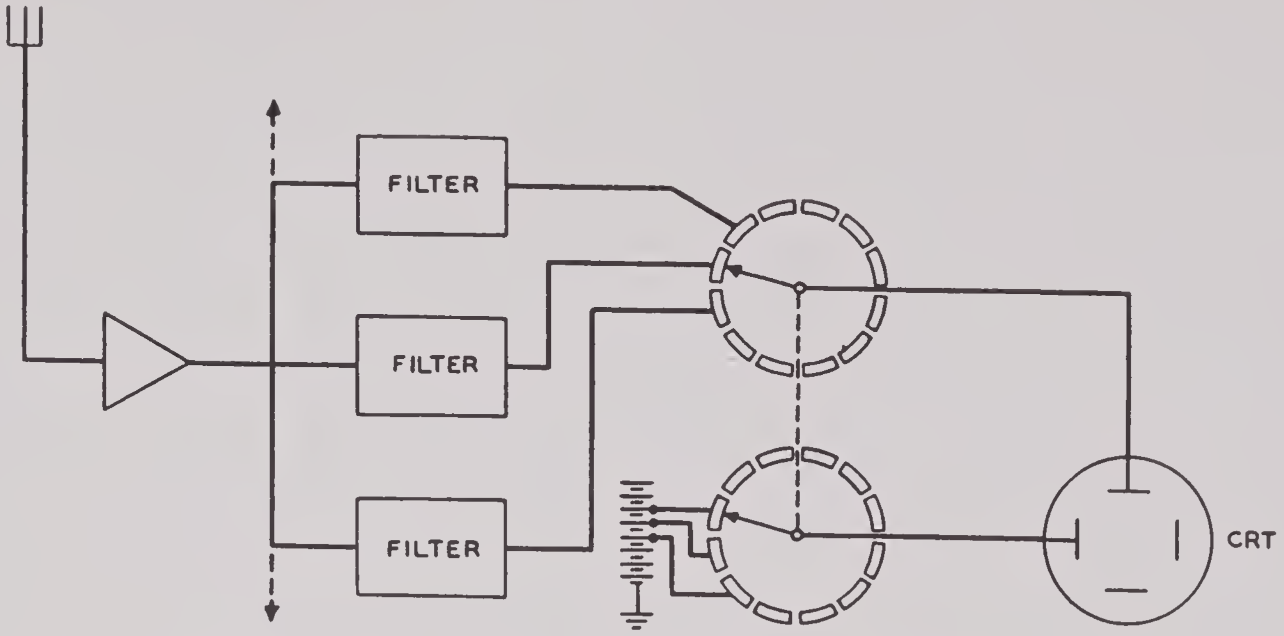

FIGURE 22. Multifilter system without frequency sweep.

or other interference present in different parts of the spectrum.

Jamming. Both broad- and narrow-band scanning are useful in studying enemy signal channels and signal characteristics and in the adjustment of jamming transmitters.

Direction Finding. Panoramic reception permits simultaneous direction finding for a number of signals of different frequency.

Radio Navigation. Panoramic receivers may be used in a variety of ways for radio navigation.[8] A simple application would be to match the amplitudes of two frequencies constituting a range beacon. More elaborate schemes include the possibility of beacons varied simultaneously in azimuth and frequency with a panoramic indicator to show the azimuth of the beacon. By suitable proportioning of the azimuth-frequency characteristic, navigation courses for moving craft can be charted in vertical, horizontal, or inclined planes.

Instrument Landing. A panoramic receiver might perform the functions of runway localizing, runway marking, and glide-path indication.[8]

Transmission Studies. Panoramic reception may be used for studies of frequency spectra, radio noise, fading, etc.

Chapter 13

PANORAMIC RECEIVER WITH MOVING-SCREEN INDICATOR

Receiver utilizing a special cathode-ray tube in which the screen is rotated so that the information it conveys can be spread out in a three-dimensional pattern, with frequency shown vertically, time horizontally, and intensity by brilliance of the trace.

13.1 INTRODUCTION

When this project[a] was started there was a great deal of dissatisfaction with the panoramic reception methods then available and NDRC was asked (especially by the Navy) to investigate new methods of panoramic presentation. In the existing systems, the presentation was in the form of a rectangular coordinate pattern with frequency shown horizontally and intensity vertically. Radio signals on different frequencies appeared as vertical pips along the horizontal frequency scale. A serious limitation was the difficulty in detecting brief radio signals.

The purpose of Project C-27 was to apply to the panoramic problem a new type of cathode-ray tube with a moving screen.[1] In this manner the information could be spread out in three dimensions with frequency shown vertically, time horizontally, and intensity by trace brilliance. Thus a telegraph signal would appear as an actual dot-and-dash trace across the screen and a-m telephone signals as a variable-intensity tone. The whole pattern of radio signals within a certain frequency interval and space of time would be spread out like the pattern on a strip of fabric coming off a loom. Brief signals would be seen as small details in this pattern and different kinds of signals should be easily recognizable.

A model was built, all parts of which were developed specially under this project except the cathode-ray tube, work on which had begun for other purposes.

13.2 RESULTS OBTAINED

In the diagrams observed on the screen, different types of signals were clearly discernible. For telegraph stations sending 40 words per minute or less, the code was clearly reproduced and could be read from the screen. Fast machine-sent telegraph signals could be recognized as telegraph but could not be deciphered. Under good conditions it was possible to observe, in the case of speech-modulated signals, a trace having serrated edges indicative of the varying frequency limits of the double side band and light and dark striations corresponding to the syllabic modulation.

The frequency resolution was such as to permit from 60 to 80 different stations with uniform spacing to be distinguished on the screen. By sweeping a narrower band, signals very close together in frequency could be spread apart.

Thus, the moving-screen receiver overcame much of the difficulty experienced with existing conventional receivers where the picture on the screen was an instantaneous diagram of amplitude versus frequency, in which the vertical spikes representing different radio stations constantly move up and down because of modulation, fading, noise, etc. Such a diagram is very trying to monitor and signals of short duration are difficult to detect.

13.3 GENERAL PRINCIPLES

A block schematic of the receiver is shown in Figure 1. Signals received from the antenna in the range 2 to 10 mc are separated by means of a band filter from signals in other parts of the spectrum. After two stages of r-f amplification, they are heterodyned to an intermediate frequency of 40 mc by means of a sweep oscillator which is varied by a reactance tube. Unwanted products of the first modulator are eliminated by a filter having a pass band of 40 mc ±100 kc. The signals are then heterodyned with a fixed carrier of 40.548 mc to bring them to the second intermediate frequency of 548 kc where they are selected by a band filter. Four different widths of band filter, 1, 5, 10, and 25 kc, are available on a switching basis. The signals then pass through a high-gain amplifier which compresses a wide range of input signal amplitudes to a very narrow range in the output. After further amplification, the signals are applied to the modulating grid of the cathode-ray tube which is biased so that the beam is turned on only when signals are present.

The frequency of the first beating oscillator is varied

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[a] Project C-27, Contract OEMsr-49, Bell Telephone Laboratories, Inc., Western Electric Co., Inc.

[redacted]

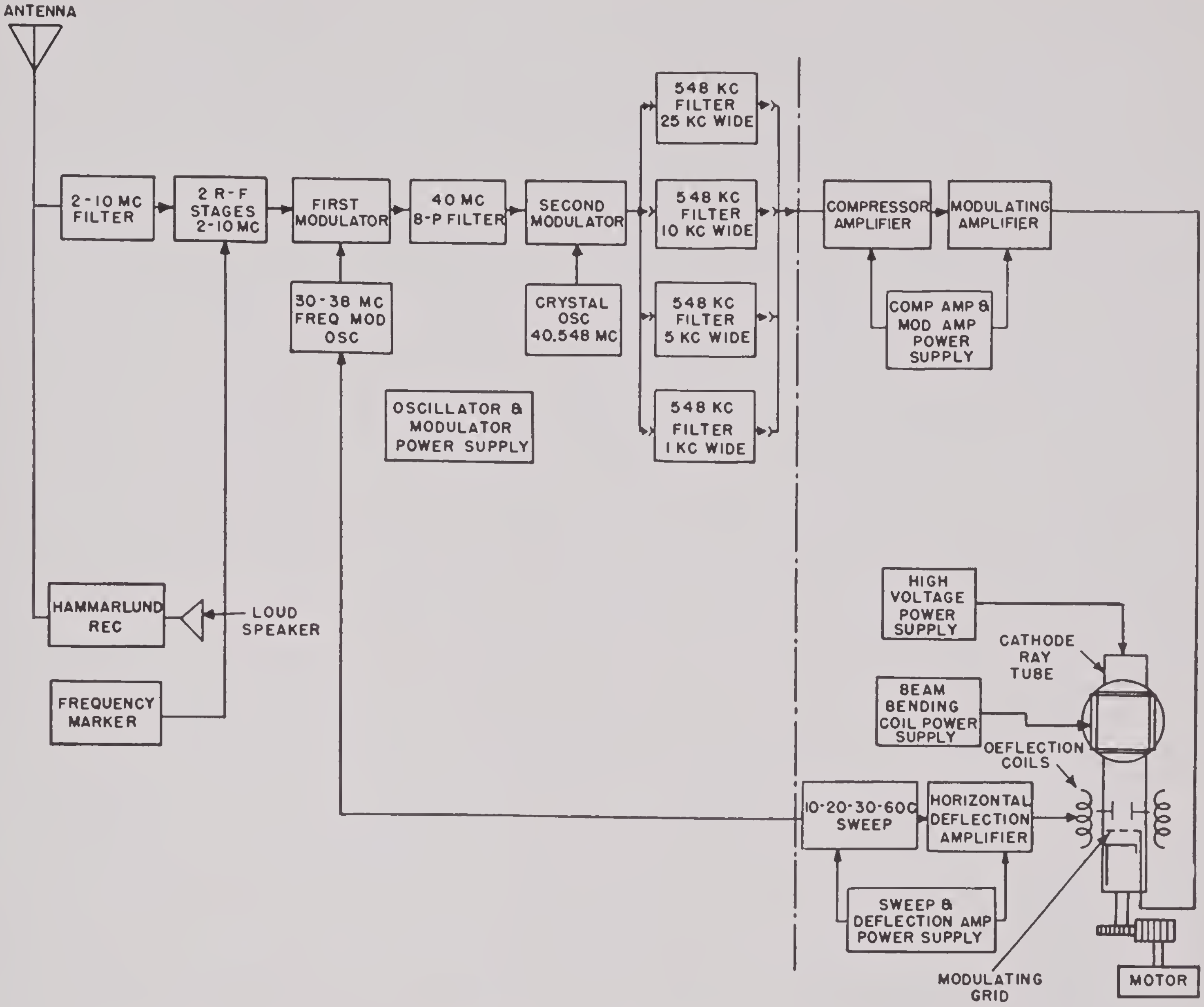

FIGURE 1. Block diagram of 2- to 10-mc panoramic scanning receiver with moving screen indicator.

by means of a reactance tube control and sawtooth generator, with provision for setting the sawtooth frequency at 10, 20, 30, or 60 per second. In addition to regulating the f-m sweep, the sawtooth wave is also applied to a pair of coils which deflect the cathode-ray beam in a vertical direction across the screen.

Rotation of the tube gives a diagram in which time is the horizontal dimension. Traces corresponding to the incoming signals are produced by persistence of the fluorescent screen. These traces appear at different vertical positions according to the frequency of the radio stations.

In the model receiver the range swept over by the f-m oscillator could be adjusted in steps from 8 mc down to 0.05 mc and arrangements were available for positioning the sweep range at different locations in the 2- to 10-mc range.

A Hammarlund 200-SPR receiver was incorporated in the model to permit aural reception of any train of signals appearing on the screen. In conjunction with this receiver there was provided an arrangement for producing on the screen a marking trace of known frequency. Thus the Hammarlund receiver could be quickly tuned to any desired trace and the frequency of that trace read on the receiver dial.

13.4 DESCRIPTION OF COMPONENTS

In the following paragraphs the individual components making up the model receiver will be described, together with their characteristics.

INPUT ATTENUATOR

To take care of extremely high signal strengths at the input, a two-step π-type attenuator designed for a 75-ohm unbalanced circuit provided losses of 0, 20, or 40 db.

2- to 10-mc Filter

This is an electric filter designed to provide not less than 35 db attenuation below 1,600 kc and above 12,000 kc, with a mid-band loss of 0.3 db and less than 1.5-db distortion over the band. The filter operates between 75-ohm unbalanced impedances.

R-F and Modulator Circuits

The r-f amplifier consists of two broad-band coupled stages using 6AC7 tubes. The output of the amplifier is applied to the control grid of the first modulator, which also uses a 6AC7 tube. The output of the first modulator is applied to the 40-mc filter, thence to the grid of the second modulator tube, also a 6AC7. This modulator operates into a balanced impedance of 125 ohms.

A buffer amplifier isolates the first modulator from its carrier oscillator. The circuit gain from the 2- to 10-mc filter input to the second modulator output is 17 ± 0.3 db.

40-mc Band Filter

To meet the requirements for this filter, two different models were built, the second of which is somewhat superior. The mid-band insertion loss at 40 mc is less than 10 db and the distortion over the pass band of 39.9 to 40.1 mc is less than 5 db. Discrimination at frequencies of 40 ± 0.3 mc is greater than 20 db and increases for frequencies farther from the pass band.

The filter consists of small mica and air condensers and lengths of coaxial line which furnish the high-Q inductances required for the narrow pass band. Through an impedance transformation, the inductance values and hence the lengths of the line sections are made identical.

First Oscillator

The first oscillator delivers a carrier frequency which varies in accordance with the amplitude of the input sawtooth voltage. It consists essentially of a Hartley oscillator which is frequency-modulated by a reactance tube. Figure 2 shows the fundamental principle involved. The grid of the control tube T_1 is supplied with a voltage E_{RF} which leads the tank voltage E_T by 90 degrees. This voltage produces in the plate circuit a current I_1 which leads the tank circuit voltage E_T by approximately 90 degrees. Hence the control tube acts like a capacitor shunting the oscillator tank circuit. The capacitance depends on the plate current I_1, which in turn is a function of the transconductance of the control tube. Thus by varying the control voltage E_C the frequency is varied.

The input control voltage consists of two components whose amplitude can be varied independently. One of these is a d-c voltage obtained from a well-regulated power supply. The other is a sawtooth voltage obtained from the sweep amplifier. Adjustable attenuators make it possible to change the width of the frequency band swept over and the mid-frequency position of this band. The frequency band covered can be adjusted to nominal values of 8, 4, 2, 1, 0.5, 0.2, 0.1, or 0.05 mc, and these bands can be positioned anywhere in the range 30 to 38 mc.

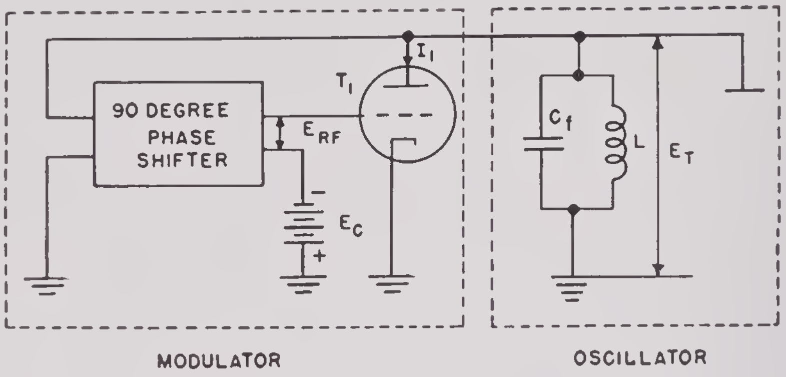

Figure 2. Reactance tube modulator used as sweep oscillator.

The diagram of Figure 3 illustrates the operation of the f-m oscillator. The curve of frequency versus control voltage approximates a straight line. Thus if the input control voltage consists of a d-c component of 6 volts and a sawtooth component of 4 volts, then the frequency of the oscillator is swept through a band having a width of approximately 2 mc and a mid-frequency of 35 mc.

The oscillator has a self-contained buffer amplifier which delivers approximately 1.5 volts rms to the grid of the buffer amplifier. Over the frequency range from 30 to 38 mc the output level is flat to within about 1 db.

Second Oscillator

This oscillator employs two 6AG7 tubes. The frequency is controlled by a bridge which is balanced for all frequencies except that of a 40.548-mc crystal, so that feedback for oscillation is available at that frequency. A buffer amplifier delivers about 2 volts to the grid of the second modulator tube.

548-kc Band-Pass Filters

By means of a switch, selection may be made of any one of four filters having band widths of 1, 5, 10, and 25 kc, respectively, all centered at 548 kc. The two narrowest filters employ crystal elements, while

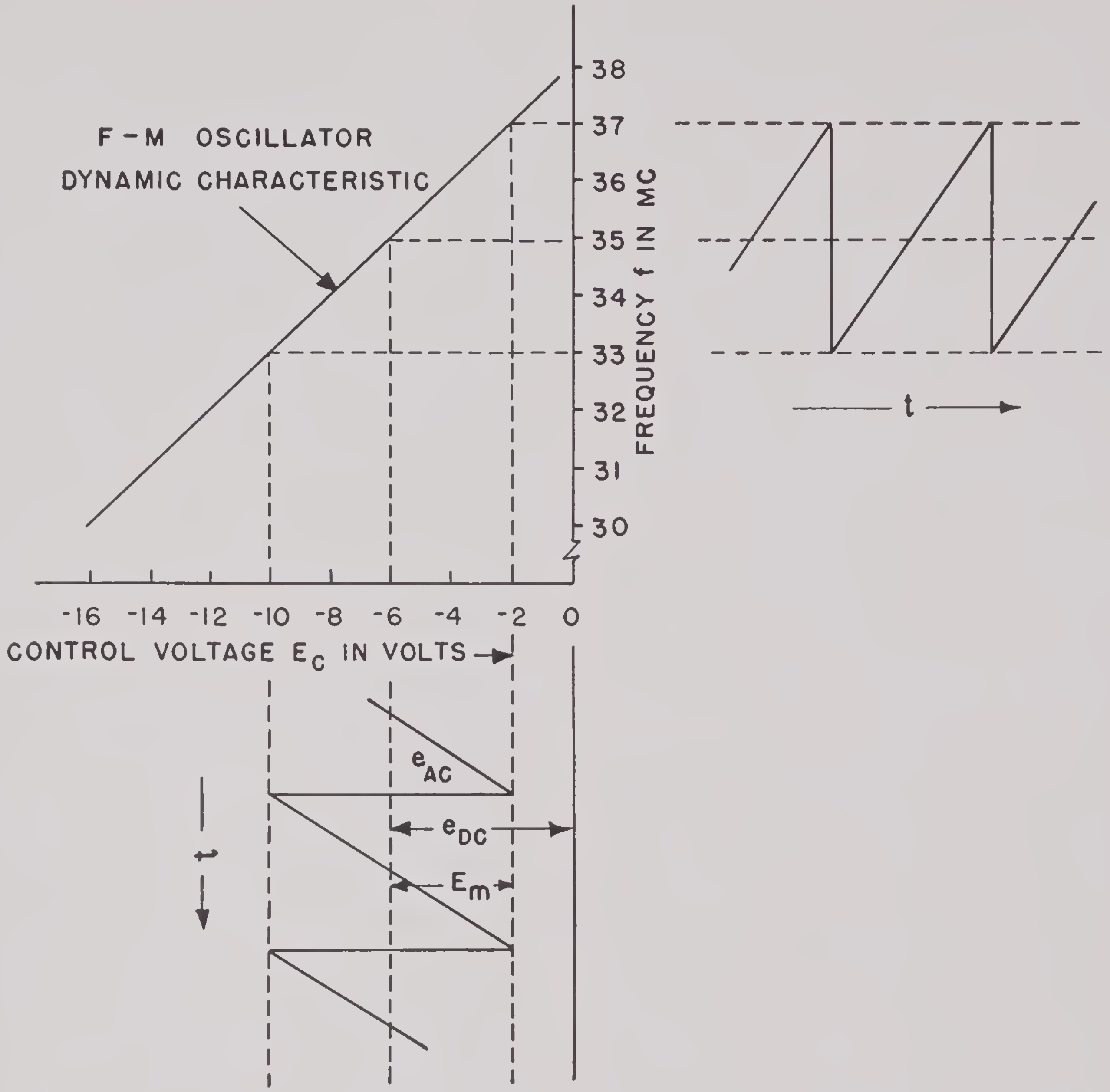

FIGURE 3. Operation of sweep oscillator.

the other two use electric elements only. The terminating impedance of the filters is 125 ohms balanced to ground. Distortion over the filter band in each case is not greater than 3 db and the mid-band loss does not exceed 6 db. In the attenuating regions the loss rises rapidly to a discrimination not less than 30 db at 548 kc ± $2f_b$ where f_b = band width. A shielded unbalanced 125:125-ohm transformer is connected between the balanced filters and the unbalanced amplifier which follows.

548-KC COMPRESSION AMPLIFIER

The compression amplifier consists of three stages made up of a 6AC7 (or a 6SJ7) and a 6H6 rectifier. It receives signals from the 548-kc band filter which may vary in amplitude from 86 to 26 db below 1 volt. Its output is delivered across a 200-ohm load to the modulating amplifier.

This amplifier has a maximum gain of 80 db. With maximum compression, an input level range of 60 db may be compressed to an output level range of 4 db. Both the gain and the compression are variable, the gain over a range of about 20 db and the compression over a range of about 60 db.

Band-pass interstage impedances for the first three stages are about 25,000 ohms at low levels. When the level across each interstage increases to a value that makes the biased diodes conduct, the interstage impedance drops to about 1,000 ohms (the conductance of the diodes). The limiting action produced by the diodes successfully cutting in as the level across the interstage increases will be understood from reference to Figure 4. The levels shown in Figure 5 are for a typical setting of gain and compression.

The compression is instantaneous. Whenever the voltage across any interstage exceeds the diode cutoff voltage, the wave can no longer rise in amplitude. Both positive and negative halves of the wave are

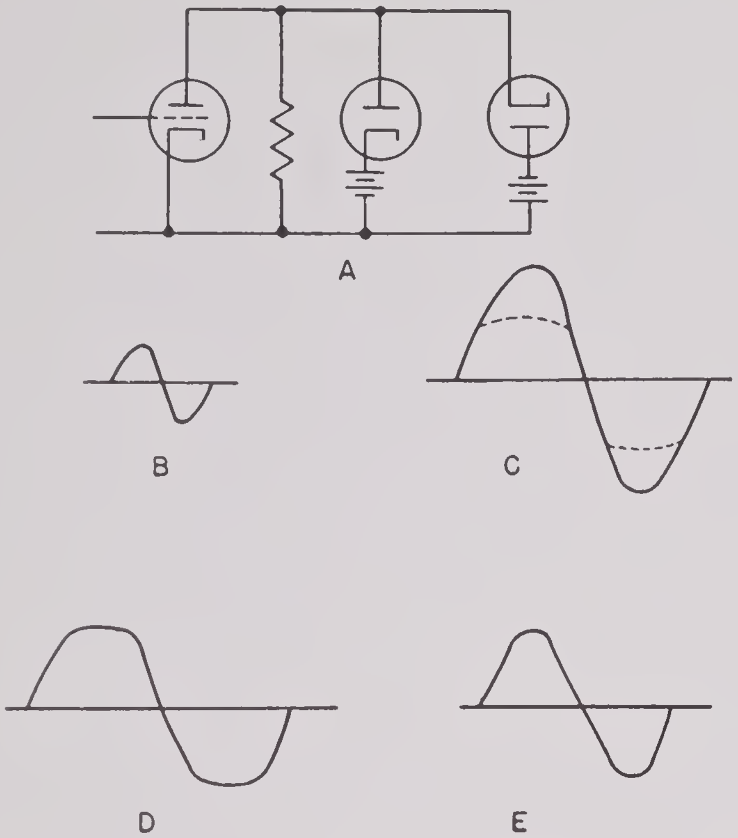

Figure 4. A, simplified schematic of clipping stage; B, input wave; C, output wave without clipping; D, output with clipping; E, output after clipping and amplification through band-pass amplifier.

clipped by the double-diode arrangement. After being distorted by clipping, the wave becomes sinusoidal again when it is amplified by following band-pass amplifier stages.

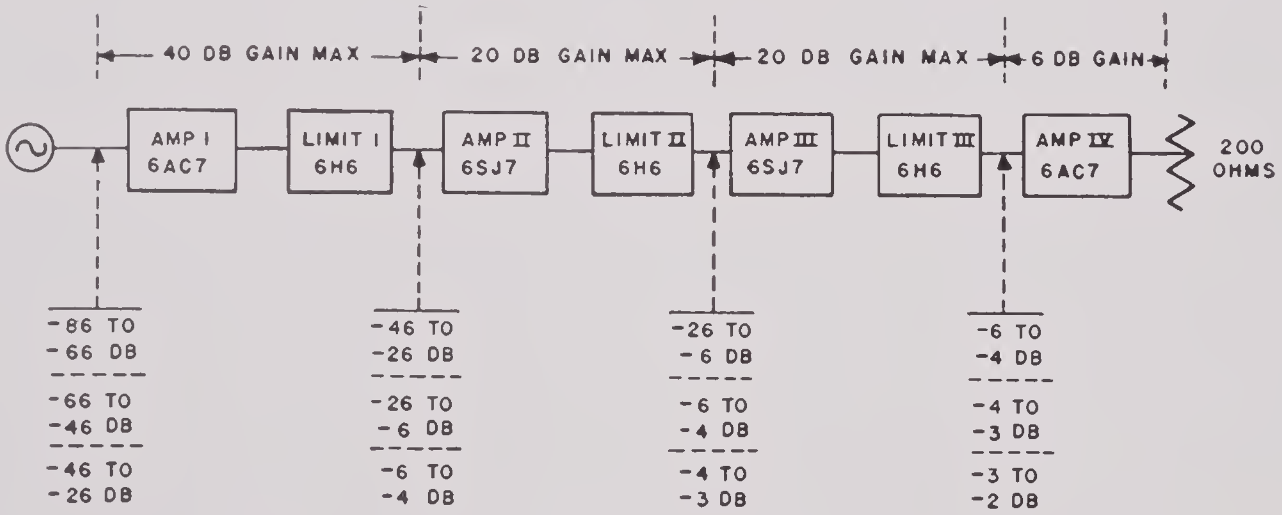

Figure 5. Operation of compression amplifier in schematic form.

MODULATING AMPLIFIER

The modulating amplifier increases the output of the compression amplifier to a value sufficient to completely modulate the electron beam of the cathode-ray tube. It receives from the compression amplifier approximately 0.75 volt peak across an input impedance of 0.5 megohm. A grid bias of approximately — 90 volts is required to completely cut off beam current in the cathode-ray tube. The bias is normally adjusted to approximately this value by the intensity control. The modulating amplifier is resistance-capacity coupled to the cathode-ray grid and its action varies the grid voltage below and above the cutoff value. To provide ample margin for discrimination against noise, the modulating amplifier is designed to produce approximately 140 volts peak.

Two high-gain pentode stages are used. The first is an inductance - compensated resistance - capacity coupled stage using a 6SJ7 tube. The second stage employs a 6AG7 high-G_m beam tube with a broadly tuned resonant output circuit which is essentially flat over the widest band received from the 548-kc filter. Cathode and screen by-passing and the interstages are designed to reduce the gain sharply outside the pass band.

CATHODE-RAY TUBE

The appearance of the cathode-ray tube is shown in Figure 6. The electron beam is emitted vertically from a gun in the lower neck of the tube, passing a control grid which regulates the intensity of the beam. The spherical part of the tube is girdled by a fluorescent horizontal band 5 in. wide. The upper and lower portions of the sphere are covered by Aquadag to which is connected the anode potential of 5,000 volts. By means of beam-bending coils a stationary horizontal electromagnetic field bends the electron beam into an approximately horizontal direction so that it falls on the fluorescent screen. A pair of deflection coils arranged in the form of a yoke is mounted around the upper end of the lower neck of the tube. When a sawtooth wave with a frequency of 10, 20, 30, or 60 cycles is passed through the deflection coils, the beam is caused to swing back and forth in the tube neck. This results in a vertical trace on the fluorescent screen.

The operating potentials are delivered to the tube through slip rings.

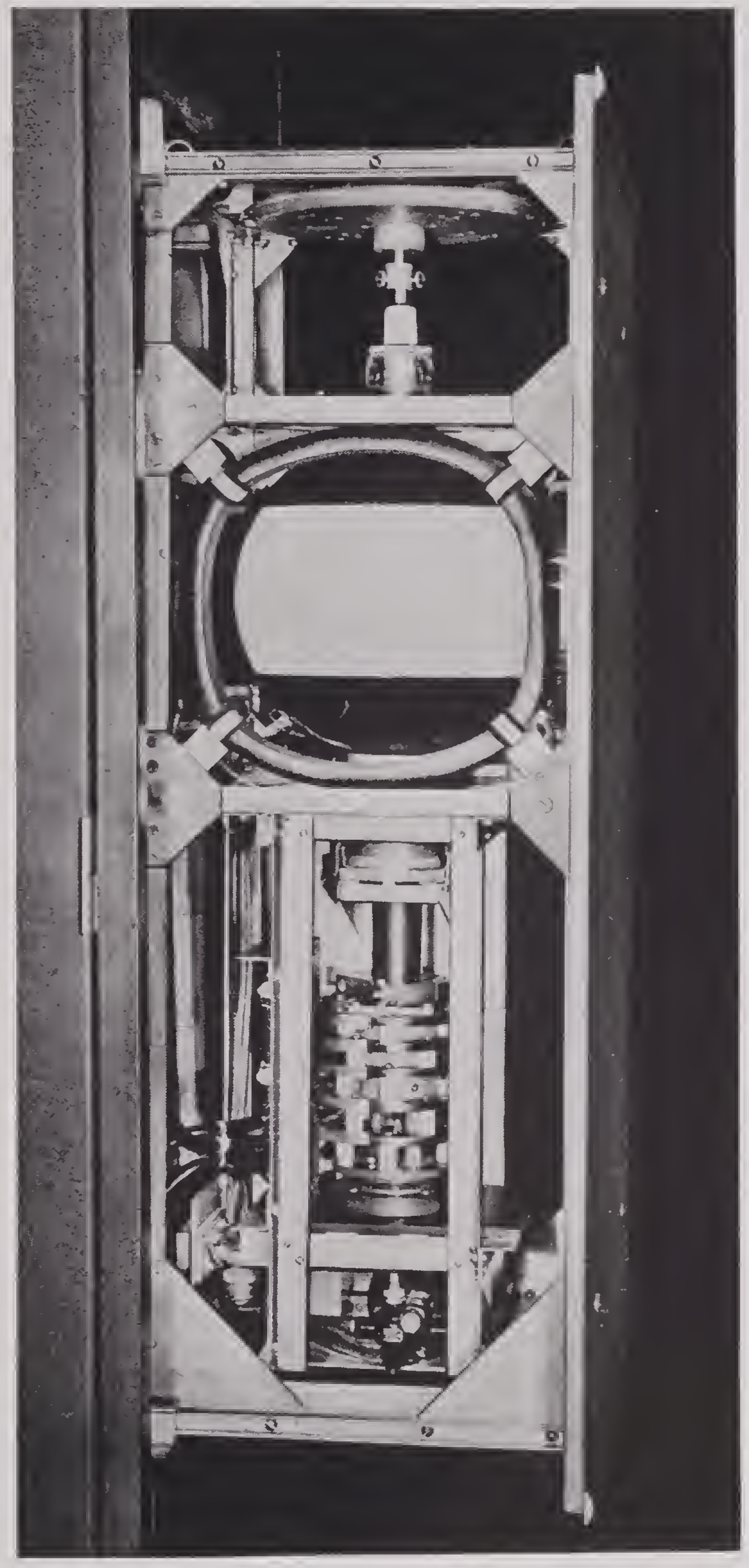

Figure 6. View of moving screen receiver showing respective positions of cathode-ray tube and screen.

Since the beam-bending coils and the deflection coils are held stationary, the electron beam always lies in a fixed vertical plane while the tube is rotated. Thus a continuously changing screen surface is provided to serve as a time axis. Since the deflection of the beam is in synchronism with the sweep of the incoming signals, a spot is obtained at each point in the vertical plane of the screen for which a signal occurs at that instant in the corresponding position in the frequency spectrum. The next trace of the sawtooth oscillator finds a fresh surface on the screen, while the original mark persists.

The tube has a diameter of 10 in. It is rotated at a rate of 1 revolution in 30 seconds, thus producing approximately 1 in. of horizontal movement of the screen per second. The persistence of the screen is sufficient so that the decay during the 5 seconds corresponding to the viewing aperture is not marked. While observations have been made in a very dark room of signals persisting on the screen for as many as 5 to 10 revolutions, i.e., 2½ to 5 minutes, the decay characteristic is such that signals coming around after one revolution of the tube do not interfere appreciably with the observance of the signal diagram.

Sawtooth Generator and Associated Amplifiers

The sawtooth generator produces a sawtooth wave of 10, 20, 30, or 60 cycles which is amplified and applied through the deflecting coils of the cathode-ray tube. The sawtooth voltage is also used to control the frequency of the first carrier oscillator.

The sawtooth generator employs two 6AC7 tubes in a direct-coupled trigger-tube circuit. One tube charges a parallel RC circuit which discharges. When it has discharged sufficiently it actuates the trigger tube which delivers a sharp pulse to the charging tube which brings the condenser back to its original voltage. The constants of the RC circuit are chosen so as to provide constant voltage output and maximum sweep linearity. A small amount of 60-cycle voltage is injected into the trigger-tube circuit to synchronize the sweep with the power-line frequency.

Adjustment of sweep rate is accomplished by a switch. A fine frequency control adjusts the RC discharge circuit to synchronization.

The output of the sawtooth generator is applied to a buffer amplifier which consists of a single 6SJ7 stage. This amplifies the sweep output voltage of five volts to approximately 20 volts peak to peak.

The buffer amplifier energizes a 6AG7 cathode follower which delivers 15 volts peak to peak to a 1,000-ohm load which supplies the sawtooth wave to control the first carrier oscillator.

The output of the buffer amplifier also energizes the current amplifier which supplies the sawtooth wave to the deflection coils. A current swing of 100 ma is required for full sweep across the screen of the cathode-ray tube. The deflection coils are direct coupled to the 6L6 output stage, the d-c component being balanced out. The plate load consists of inductance and resistance whose time constant is made equal to that of the deflection coils. Vernier controls are provided to adjust deflection yoke bias and the time constant of the plate circuit.

Marking Arrangement

In the marking arrangement used on the model, the heterodyne frequency provided by a Hammarlund 200-SPR receiver, which is 465 kc above the frequency to which the receiver is tuned, is modulated with a carrier of 465 kc. The output of the modulator is applied to the input of the 2- to 10-mc filter, thus producing on the screen three marking frequencies, the lowest one of which is the one to which the Hammarlund receiver is tuned. When this lowest marking line is made to coincide on the screen with any observed signal train, the frequency of that train is determined and the signals are received aurally. A different marking circuit was developed subsequently and is described below.

Power Supplies

All the 300-volt anode supplies are closely regulated to compensate changes in input voltage and variations in the load. Over a range of 105 to 125 volts a-c input and a 30 per cent range of output current, the regulation is within ±0.05 volt. Over the usual working range of input voltage and load, the regulation is of the order of ±0.02 volt. The circuits consist of the usual type of rectifier with a capacitance input filter. The unusual degree of regulation is achieved by a special feedback amplifier working into two or more paralleled current valves or control tubes, the number depending on the requirements. Any small change in output voltage produced by a change in input voltage is amplified by the 6SF5 feedback amplifier tubes shown in Figure 7. The resulting amplified variation applied to the grids of the 6Y6G current valves varies their impedance, thus compensating the voltage change in the output.

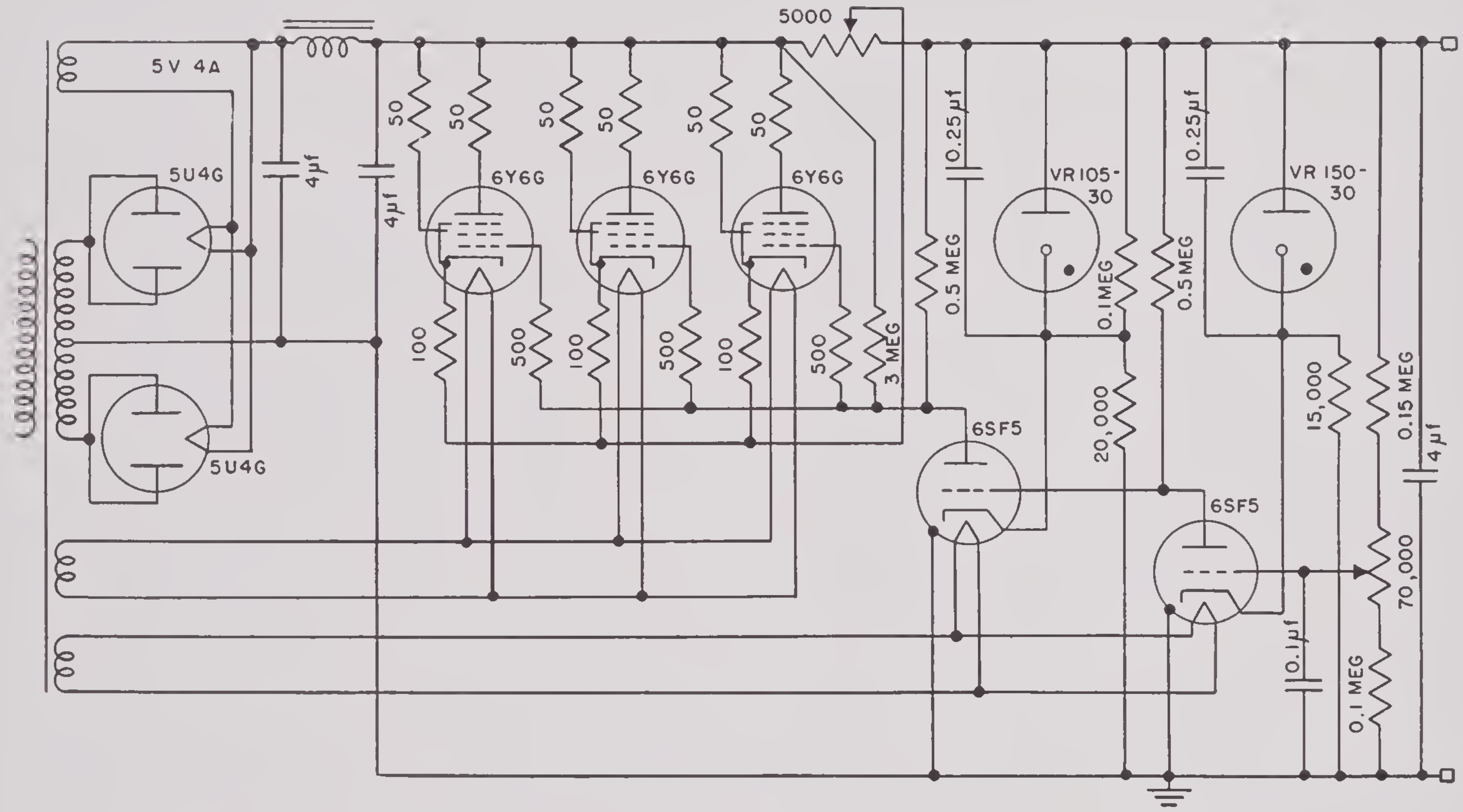

Figure 7. Circuit diagram of voltage-regulated power supply for 250-volt, 300-ma output.

13.5 RESEARCH WORK

Scanning Oscillator

Considerable research was required to develop the scanning oscillator, since no previous knowledge was available regarding the design of an electronically controlled oscillator with such a wide percentage sweep at such a high frequency. Extremely meticulous design was required to obtain satisfactory stability.

In the course of work on the scanning-oscillator problem an alternative involving the use of an RC oscillator was studied. In this the frequency would be varied by changing the bias of a diode to vary the resistance of the oscillator circuit. It appeared that an oscillator of this type might be used at a frequency higher than previous RC oscillators but in view of the constants involved in operation at 30 to 38 mc it was planned to use a lower fundamental frequency swing in the range 15 to 20 mc and select out the second harmonic for the desired variable carrier. A circuit was designed which appeared capable of providing a wider percentage frequency swing than the reactance-controlled oscillator. Difficulties were experienced, however, in obtaining stability, and although it appeared that these could be solved this approach was abandoned when the reactance-tube scheme proved successful.

Some study was also made of the application of feedback to the present sweep oscillator but this was not carried to completion.

Compression Amplifier

Study was required of different methods of handling a wide range of input signals. This indicated that the preferred method would be to use a compression amplifier. Since an amplifier of this type with such a high compression range had not previously been built, research was required to determine the best circuit design.

Infrared Wiping Out

Originally it appeared that it would be desirable to wipe out the traces on the screen after passing the viewing window, so that these traces after completing one revolution would not interfere with the new traces. Some work was done with an arrangement employing infrared rays to obtain such wiping out. The infrared was obtained from ordinary lamps. With the filtering used, the amount of light in the lower part of the visible spectrum appearing on the screen seemed to be somewhat detrimental to observation. Although it appeared that a satisfactory wiping-out arrangement could be secured, further work was abandoned when it was found that the traces persisting after one revolution did not interfere appreciably with observation.

13.6 FURTHER WORK

After demonstrating the preliminary model, certain improvements and modifications were suggested, so that studies of microsecond pulses in the frequency range 500 to 600 mc could be made.

The frequency-marking arrangement, whereby the receiver is tuned to any signal appearing on the screen, was modified so that only one marking frequency appears on the screen. The new arrangement employs an additional tuning unit of the same type as that used in the Hammarlund receiver. This unit is modified so as to produce a frequency which differs by 465 kc from the heterodyne frequency of the receiver and therefore is the same as the frequency to which the receiver is tuned. The marking unit is coupled mechanically to the receiver by means of a sprocket and chain which drives the band-changing switch and by a phosphor bronze cable and pulley which drive the tuning condenser. In this way, fairly close tracking is obtained. With the marking frequency in operation, a trace moves across the screen as the tuning dial of the receiver is turned. When this trace is superposed on a signal trace, that signal is heard. The frequency of the station may then be determined from the tuning dial. Evidently it would be possible to connect a transmitter to the frequency marker and thus jam the signals in question.

Study of Scanning Problem

During the development work, certain fundamental investigations were necessary. For example, in a scanning receiver the build-up time of the selecting filter limits the rate at which the given frequency band can be scanned, or, if the scanning rate is fixed, limits the width of frequency band covered. Conversely, if the frequency band and scanning rate are fixed, then the width of the selecting filter and the consequent degree of resolution are limited. This problem had been studied by various investigators prior to the inception of this project. (See Chapter 12.) To make it possible to study this relationship as applied to the moving-screen type of scanning receiver and to permit different degrees of resolution of the frequency band as desired, provision was incorporated in the experimental model for (a) different widths of selecting filter, namely 1, 5, 10, and 25 kc, (b) different scanning rates, namely 10, 20, 30, and 60 cycles per second, and (c) different frequency ranges from 8 mc to 0.05 mc. Diagrams employing sweep rates of 30 cycles or less have been found to be unsatisfactory due to the noticeable gaps between successive marks for different scanning cycles. Accordingly the 60-cycle sweep is generally to be preferred. With this sweep the entire 8-mc band can be covered using a 10-kc filter. For narrower bands greater resolution can be obtained with a narrower filter.

Pulse Reception

Preliminary study of pulse reception in the range of several hundred megacycles indicates that with pulses which are on for only a very small percentage of the time and with a total signaling period which may be very brief, the method of scanning the frequency range as employed in the present receiver may be undesirable because the signals might be missed. On the other hand, because the pulses are extremely short compared to the period between them, there is a small probability that pulses from more than one station will occur simultaneously. Hence it is believed possible to receive a broad band of frequencies and to sort out the pulses which are found. Such sorting out might be accomplished by passing the received signal band through a limiter which reduces all pulses to the same amplitude and then through a discriminator which makes the amplitude of the pulses proportional to their frequency. The output of the discriminator could then be applied to a moving-screen cathode-ray tube so as to produce a diagram in which there would appear for each pulse train a horizontal line whose vertical position would indicate the radio frequency and whose length would correspond to the duration of the pulse train. The general method is illustrated in Figure 8A.

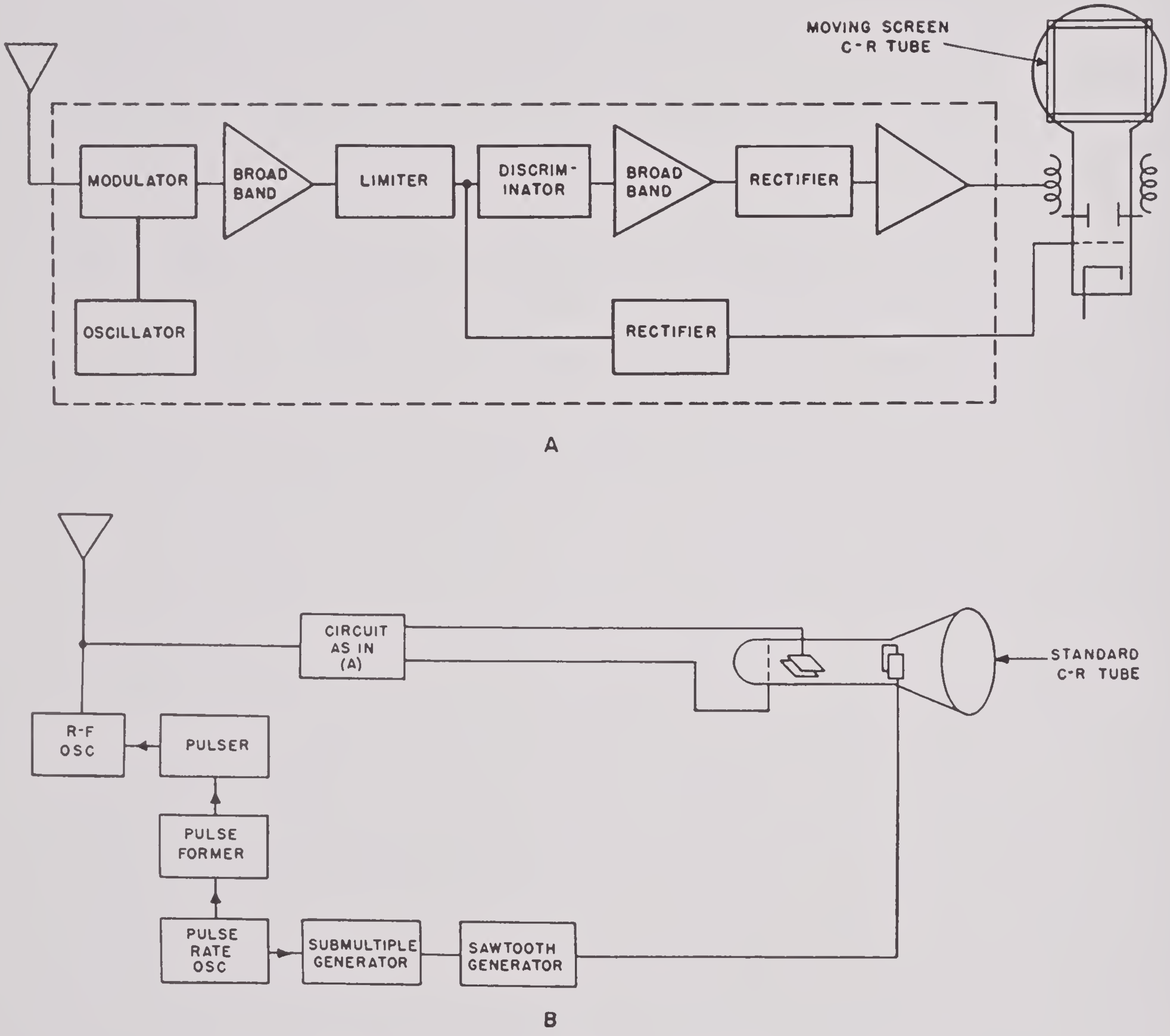

Figure 8. Proposed method of receiving short pulses on panoramic receiver.

Instead of the above, the output of the discriminator might be connected to the vertical deflecting plates of a cathode-ray tube of the usual type as indicated in Figure 8B. Thus, whenever pulses were present, the vertical position of the electron beam would correspond to the radio frequency of the transmitting station. Now if a low-frequency sawtooth wave were used as a horizontal sweep, the pulses would move across the screen at a rate sufficiently rapid to form continuous lines. With flat-topped pulses, these lines would stand out in comparison with the traces produced by the sides of the pulses. Moreover, by applying the constant-amplitude pulses in the output of the limiter to the modulating grid of the tube, the traces during the intervals corresponding to the sides of the pulses could be largely blanked out. Thus the device would detect the presence of pulses and indicate their radio frequency but not the duration of the pulse train. If the sawtooth frequency were varied until a frequency is reached which is a submultiple of the pulse frequency of a particular pulse train, those pulses would be stopped on the screen and the pulse rate could be determined. With either arrangement shown in Figure 8 it might be desirable to introduce a marking frequency as a simple means of frequency determination.

Chapter 14

IMPROVED PANORAMIC RECEIVER

Further studies on panoramic reception, development of an improved long-scale model for the 3- to 10-mc range, study of multiline indicators, methods of indicating marker frequency.

14.1 INTRODUCTION

Under Project C-27 a panoramic receiver was constructed which scanned the approximate frequency range of 3 to 10 mc or any of various narrower bands located anywhere in that range. The indicator employed was a special cathode-ray tube rotated to give a continuous motion of the screen in relation to the scanning beam. The sawtooth wave which controlled the f-m sweep deflected the beam in the cathode-ray tube in a vertical direction representing frequency while rotation of the tube gave a diagram in which time was the horizontal dimension. The equipment was built for laboratory tests and, therefore, had many more variables that would be useful in a field receiver.

Tests indicated that the length of the frequency scale, approximately 4 in., was a limitation, particularly when scanning a wide band in which a large number of signals were present. It was also desirable to extend the frequency range from 10 to 30 mc. Numerous other factors needed further investigation. Accordingly, under Project C-36[a] a long-scale 3- to 10-mc model[1] was constructed, studies were made on scanning filter design, multiline indicators were examined, methods of indicating marker frequency were developed, and techniques were worked out for panoramic reception in the range from 0.1 to 30 mc.

14.2 LONG-SCALE 3- TO 10-MC RECEIVER

Before actual work could proceed on the lower-frequency receiver built under this project, several factors had to be determined. They are outlined in the paragraphs which follow.

14.2.1 Length of Scale

In studying methods of providing a long-scale indicator, theoretical and experimental investigations were made on length of scale, type of diagram, type of indication, scanning rate, type of cathode-ray tube and other factors. As a result a diagram in the form of a spiral on a 7-in. tube was chosen.

Assuming no limitation in the resolution capabilities of the equipment, the amount of information that can be portrayed on the screen increases directly with length of scale. However, it is readily possible with available technique to display an amount of information far exceeding the capabilities of the observer. Thus for example, if radio stations were located at 5-kc intervals throughout a band of 8 mc and the scale length associated with each station were held to 0.2 in., the scale length would be greater than 300 in. Even though many stations were missing it would be quite impossible for an observer to monitor such a 300-in. scale. The length of scale was therefore determined by the capabilities of the observer rather than those of the scanning receiver.

Observer capability will vary widely with observers and with the purposes for which observations are made. However, television experience indicates that screens 5 in. square are about right for arm's length viewing of entertainment. This seems to indicate that the screen of a 7-in. cathode-ray tube is at least adequate for presenting all the panoramic receiver indications that an observer can consistently use. The length of trace may be varied at the discretion of the observer and experience indicates that 40 in. is a good value for most conditions.

14.2.2 Shape of Scale

Ease of viewing requires that the scale be compressed into an area not much wider than its height. Accordingly, the scale should be doubled up in some manner.

A retrace discontinuity during the scanning interval is avoided by bending the trace into a spiral. This also affords a substantially uniform distance between adjacent lines. The spiral may be adjusted to any desired length of trace and to occupy any desired per cent of the screen area by changing the number of turns of the spiral and adjusting the distance between turns. For normal operation the spiral in the 3- to

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[a] Project C-36, Contract No. OEMsr-357, Bell Telephone Laboratories, Inc., Western Electric Co., Inc.

10-mc model is operated with 4 turns spaced about ½ in. apart. The frequency band scanned during each turn is approximately the same. The progression of the scale is from high frequencies nearest the center to low frequencies at the outer end. Radio stations tend to have less frequency separation at the lower-frequency end of the band, so the outer or longer turns of the spiral are used for the low frequencies.

14.2.3 Method of Obtaining Spiral Trace

The electron beam of the cathode-ray tube may be caused to follow an elliptical path by applying alternating currents in quadrature phase relationship to the vertical and horizontal deflection coils. If the magnetic fields set up by the quadrature currents have the same value and are single-frequency sinusoids, the trace becomes a circle whose radius is determined by the peak magnitude of the magnetic field. If the magnitude of the alternating current in each set of coils is increased simultaneously at a rate linear with time, the trace will become spiral.

In the 3- to 10-mc receiver the output of a variable frequency oscillator creates the quadrature currents. The currents of each phase are then modulated in magnitude by the same sawtooth pulse that controls the reactance tube of the sweep oscillator circuit in the scanning receiver. Accordingly the spiral is initiated at the beginning of each sweep of the scanning receiver and each discrete frequency in the band scanned is represented by a particular position on the spiral. The sweep circuit is synchronized with a variable-frequency oscillator rather than with the power-supply frequency as in the case when the moving-screen indicator is used.

The number of turns in the spiral is determined by the ratio between the frequency of the variable-frequency oscillator (and hence of the rotating field) and that of the sweep circuit. The usual settings are 60 cycles per second for the variable frequency oscillator and 15 cycles per second for the sweep circuit. This gives a 4-turn spiral whose size is determined by the magnitude of the output of the variable frequency oscillator. The distance between turns is determined by the magnitude of the sweep power used to modulate the quadrature currents.

14.2.4 Methods of Indicating Signals

Intensity modulation, which is used in the model, results in a diagram which is pleasing to the observer and produces a minimum amount of confusion between adjacent turns of the spiral. Experience with the model indicates that there is a sacrifice in resolution due to this method of indication. How much of this sacrifice might be eliminated by a different design of the scanning filters and of the compression arrangement has not been determined. Also it is not known which is the more important, improved resolution or a more pleasing type of indication.

By swinging the electron beam at right angles to its sweep path when the signal is received, more detail and hence better resolution are obtained. This also permits more distortion in the scanning filters. However, the indicator pattern resulting from modulated signals, static, etc., dances violently. It may be that the resultant confusion and observer fatigue would outweigh the improved resolution. While sidewise deflection can be used with the spiral trace, a parallel-line diagram is better adapted for this purpose.

14.2.5 Scanning Filters

The resolution obtainable with different scanning filters at various scanning rates and band widths has been studied experimentally and compared with the optimum theoretically obtainable. An important cause of degradation is delay distortion in the filter, i.e., different transmission times for different frequencies. The signal response is prolonged due to the delay distortion and a weak signal located near a strong one may be completely masked. Another difficulty is that when a compression amplifier follows the filter, the effect of sloping cutoff of the filter is to widen the pass band as the input signal strength becomes greater. This tends to mass signals at adjacent frequencies. Unfortunately the tendency in normal filter design is for the delay distortion to increase as the sides of the attenuation characteristics are made steeper. Accordingly, obtaining optimum characteristics for the scanning filters is a very complicated matter.

Perhaps one of the most important results of this research project was that concerned with improvement of resolution by reduction of distortion in the scanning filter. Unless special attention is given to this factor in the panoramic receiver design, the resolution may suffer materially.

14.2.6 Type of Cathode-Ray Tube

Tests of long- and short-persistence screens indicated that long persistence has a definite advantage for this type of work. Accordingly an 1813-P7 tube having a cascade screen was selected. The fluorescent color was blue-white and the phosphorescent color was orange. An orange filter was provided so that the fluorescence could be eliminated if desired.

Blanking Out Recognized Stations

A simple blanking out was obtained by painting over the spots produced by identified signals. A slow-drying black paint was used so that the screen could be wiped clean at frequent intervals. (See Section 12.8.)

14.2.7 Rate of Scanning

The model is arranged for scanning rates of 4, 15, 30, and 60 per second. With the long-persistence screen and orange mask, flicker is not objectionable at the higher rates and not serious even at the 4-per-second rate. The greater resolution obtained with the lower rate is distinctly worthwhile if the desired stations are closely grouped in frequency or differ greatly in magnitude. However, the 4-per-second rate has a very definite effect on the speed with which the marker can be adjusted to an unknown signal in order that the monitoring receiver may be tuned to it for identification purposes. With the slow scan, rapid movement of the tuning control of the monitoring receiver causes the marker spot to move in jerks and become confused with spots associated with modulated signals such as telegraph signals. This is because the frequency rate of change of the marker becomes comparable to that of the scanning oscillator.

14.2.8 Reactions on Scanning-Receiver Design

The use of the long-scale indicator imposes more severe requirements on the scanning-receiver design than does the use of the moving-screen indicator.

1. The increased scale affords greater resolution and hence imperfections such as small nonlinearity in the sawtooth scanning wave become more noticeable.

2. The sweep oscillator is no longer synchronized with the 60-cycle power supply. It is now synchronized with the variable frequency oscillator used to create the spiral trace. This means that the sweep-control circuits must be carefully protected against interference from the 60-cycle power circuits. Otherwise the phasing in and out of the interference and the variable-frequency oscillator output will cause erratic sweep times and distortion of the pattern. At times, 60-cycle commercial power was used in place of the output of the variable-frequency oscillator. This is satisfactory if the commercial power is stable in voltage and frequency and has exceptionally pure wave form. Otherwise, as is the usual case, the indicator pattern is distorted and the same signal does not appear in the exact same spot during successive scans.

The scanning control circuit wiring of the original model (Project C-27) was changed to reduce coupling with the commercial power circuits. Special care in grounding arrangements is necessary in each operating location in order further to reduce 60-cycle coupling.

14.3 MODEL EQUIPMENT WITH SPIRAL-TRACE INDICATOR

The 3- to 10-mc spiral-trace indicator was used with the same h-f equipment constructed for the moving-screen indicator (Project C-27). Components shown within the dashed lines of Figure 1 comprise parts of the original equipment.

Referring to Figure 1, the operation of the h-f equipment used with the spiral-trace indicator is as follows. Incoming signals in the 3- to 10-mc range, after selection and amplification, are heterodyned to an i-f frequency of 40 mc by a beat oscillator whose frequency is controlled by a sawtooth wave through a reactance tube. The range swept over by the beating oscillator can be adjusted either to 7 mc or to any of a number of narrower bands which may be positioned anywhere in the 3- to 10-mc range. The signals from the 40-mc i-f stage are heterodyned with a fixed carrier to bring them to a second intermediate frequency of 548 kc. At this point a relatively narrow band filter is used to separate out the different incoming signals in succession. Several different widths of band filter are available for use with different rates of sweeping over the frequency band and different widths of band swept over. The signals from the 548-kc filter are passed through a high-gain amplifier which compresses a wide range of input amplitudes to a very narrow output range. The signals then pass to the long-scale indicating equipment.

14.3.1 Operation of Indicating Equipment

The operation of the long-scale indicating equipment is as follows. Signals from the compression am-

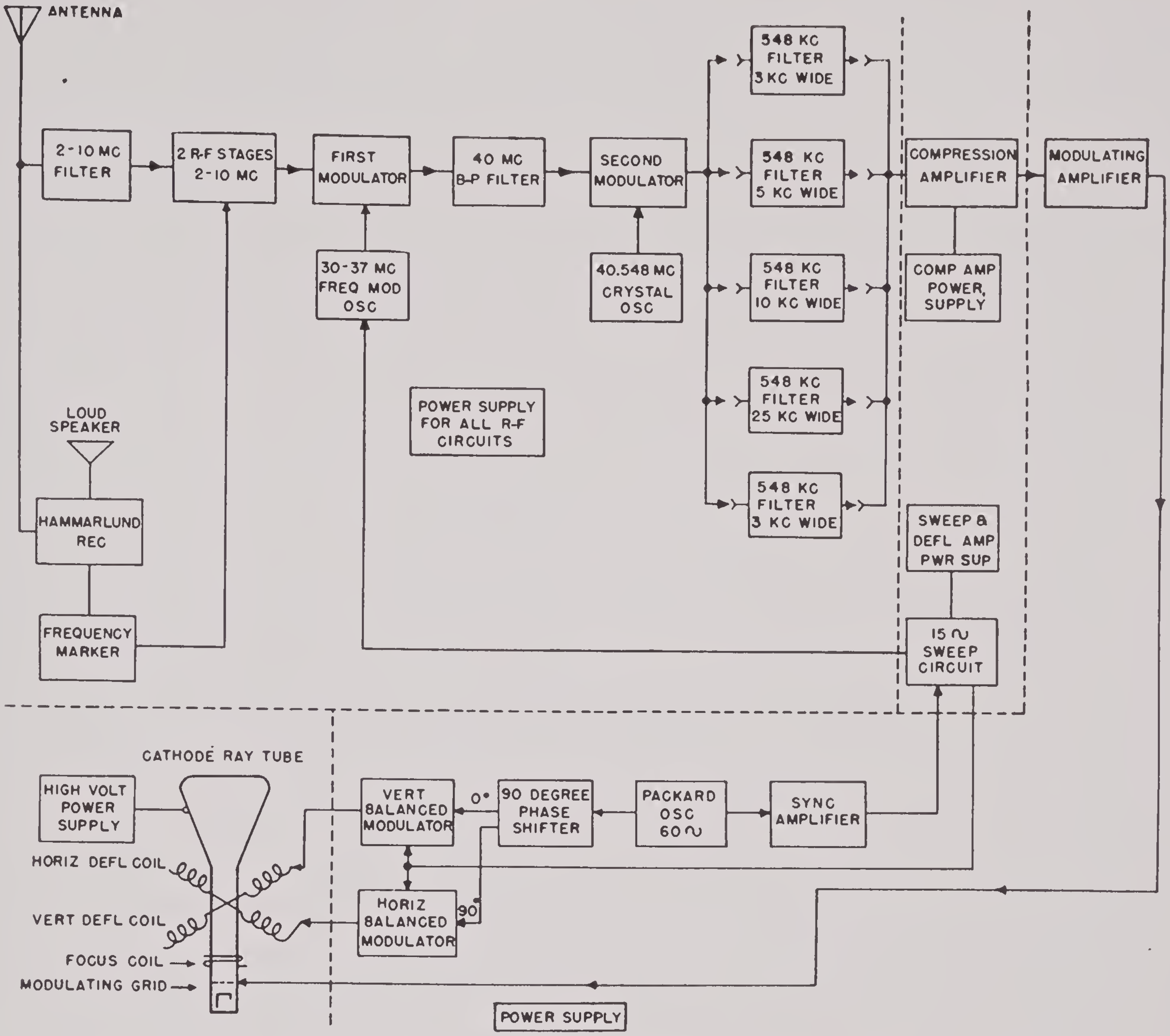

FIGURE 1. Block diagram of long-scale 3- to 10-mc spiral-trace receiver.

plifier are passed through a modulating amplifier to obtain a voltage sufficient to modulate completely the electron beam of the cathode-ray tube. The spiral trace is obtained by deriving two quadrature components of a sine wave and modulating these with a sawtooth wave whose frequency is a submultiple of the sine wave. The two resultant waves are applied to the vertical and horizontal deflecting coils, producing a spiral for which the number of turns is equal to the ratio of the sine wave frequency to the sawtooth frequency.

Each received signal is indicated by a change in the intensity of the beam, giving a bright spot on the tube. Since the same sawtooth frequency is employed both for controlling the frequency of the f-m oscillator and for producing the spiral, the frequency of the signal is indicated approximately by the position of its spot on the spiral. The highest frequency signals appear near the center and the lowest frequency signals near the outer end.

The spots for a given frequency signal are superimposed during successive scannings. This permits increased signal resolution as compared to a moving-screen indicator because the scanning rate may be lower and the screen distance between signals may be larger. The high-persistence screen reduces flicker inherent to low scanning rates.

The model is arranged so that different numbers of turns in the spiral and different sizes of spiral can be obtained. This affords a wide range of scale lengths. Experience has indicated a preference for a 4-turn spiral which affords a total scale length of about 40 in. A longer scale or larger number of turns appears to confuse the observer.

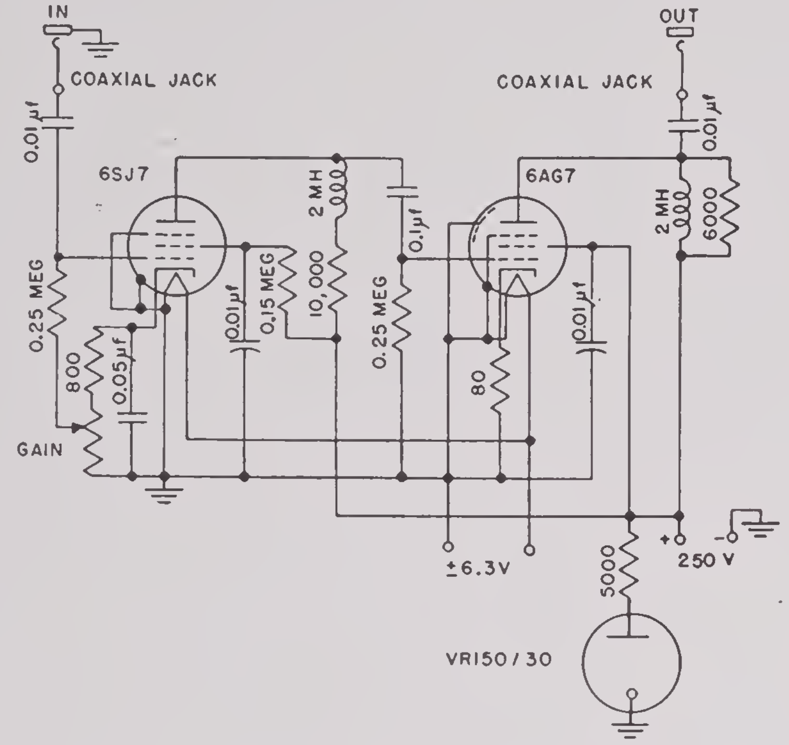

Figure 2. Circuit of amplifier for modulating beam of cathode-ray tube.

Incorporated in the model receiver was a Hammarlund receiver with which signals of any frequency within the band being scanned could be aurally monitored. In conjunction with this receiver there was provided an arrangement for producing on the screen a marking trace of the same frequency as that to which the Hammarlund was tuned as a means of determining the radio frequency of the signal being observed.

14.3.2 Modulating Amplifier

The circuit of the modulating amplifier is shown in Figure 2. This amplifier increases the output of the compression amplifier to a value sufficient to modulate the electron beam of the cathode-ray tube over the useful range of intensity.

The modulating amplifier receives from the compression amplifier approximately 0.75 volt peak. The grid bias is normally adjusted by the intensity control to a value near —50 volts, which is required for substantially complete cutoff of the beam current. The modulating amplifier varies the grid voltage below and above the cutoff value. To provide ample margin for discrimination against noise, the amplifier is designed to produce approximately 140 volts peak. The second stage has a broadly tuned resonant output circuit which is essentially flat over the widest band received from the 548-kc filter with sharp reduction in gain

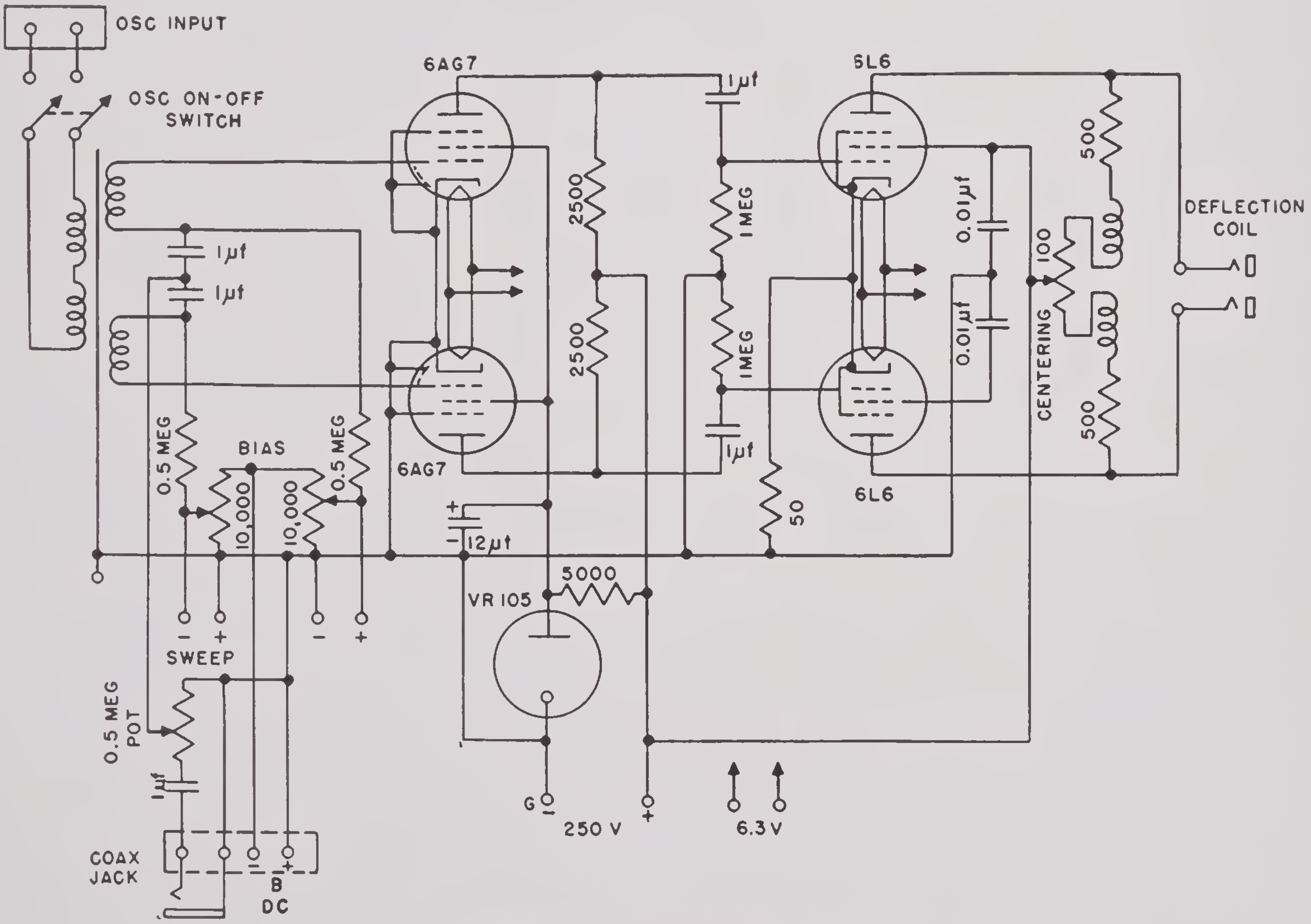

Figure 3. Circuit of balanced modulator and deflection amplifier, 3- to 10-mc receiver.

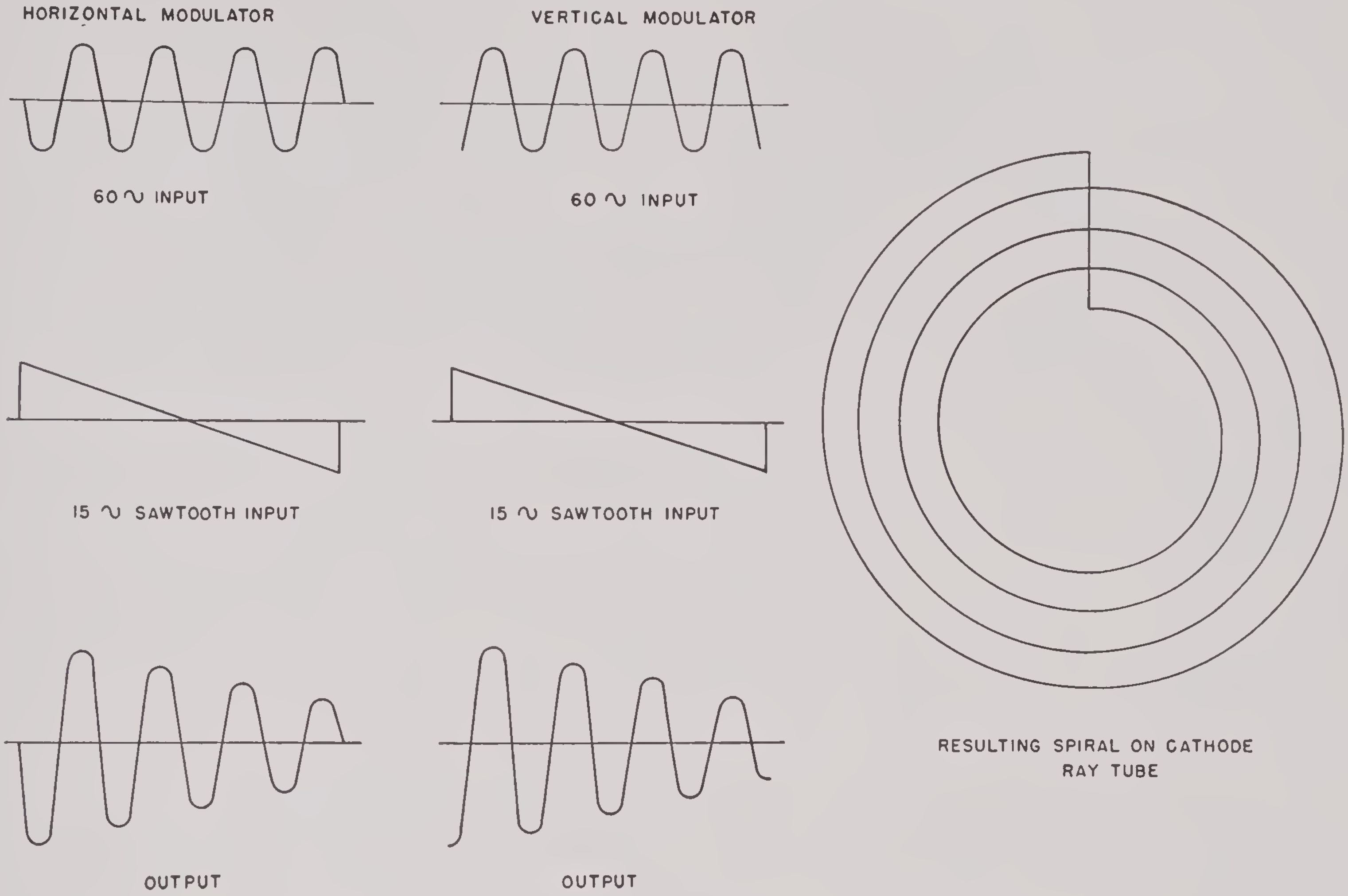

FIGURE 4. Production of spiral trace for 3- to 10-mc receiver.

outside the pass band. The gain control has approximately a 6-db range.

14.3.3 Balanced Modulators

The circuit of the balanced modulator is shown in Figure 3. This circuit consists of two push-pull stages with resistance coupling between stages. A 60-cycle sine-wave voltage is fed to the two grids of the first stage in phase opposition. A 15-cycle sawtooth voltage is fed in phase to the same grids. By means of separate bias controls on each of the input tubes, the modulator may be closely balanced so that the sawtooth wave is balanced out at the plates of the output stage where one of the deflection coils is directly coupled. The output current is then a 60-cycle sine wave, amplitude modulated by the sawtooth wave. A control on the sawtooth input voltage allows adjustment for the desired percentage of modulation. The plate load consists of inductance and resistance whose time constant is made equal to that of the deflection coil. This reduces distortion of the output wave. A potentiometer is provided in the plate load to provide for a small amount of unbalance of the d-c plate voltages, thus allowing direct current to flow through the deflection coil. This adjustment is used for centering the trace on the screen.

One of these balanced modulators was connected to the horizontal deflection coil and another to the vertical deflection coil. The only difference in the operation of the two circuits is that the 60-cycle inputs are in quadrature. This produces two output waves as shown in Figure 4. With these two waves applied to the two deflection coils, the resultant trace will be the four-looped spiral shown.

The modulators are fed with the same sawtooth wave that controls the frequency of the f-m oscillator. The output of the modulators can swing up to about 100 ma, which is required for full-scale deflection on the cathode-ray tube.

14.3.4 Phase Shifter

The circuit of the phase shifter is shown in Figure 5. This is a balanced circuit which takes a single 60-cycle sine-wave input and converts it to two outputs that are in quadrature. A variable attenuator is provided in the nonshifted output to insert loss in this branch equal to the loss of the phase-shifting network, so that the levels applied to the vertical and horizontal modulators are equal.

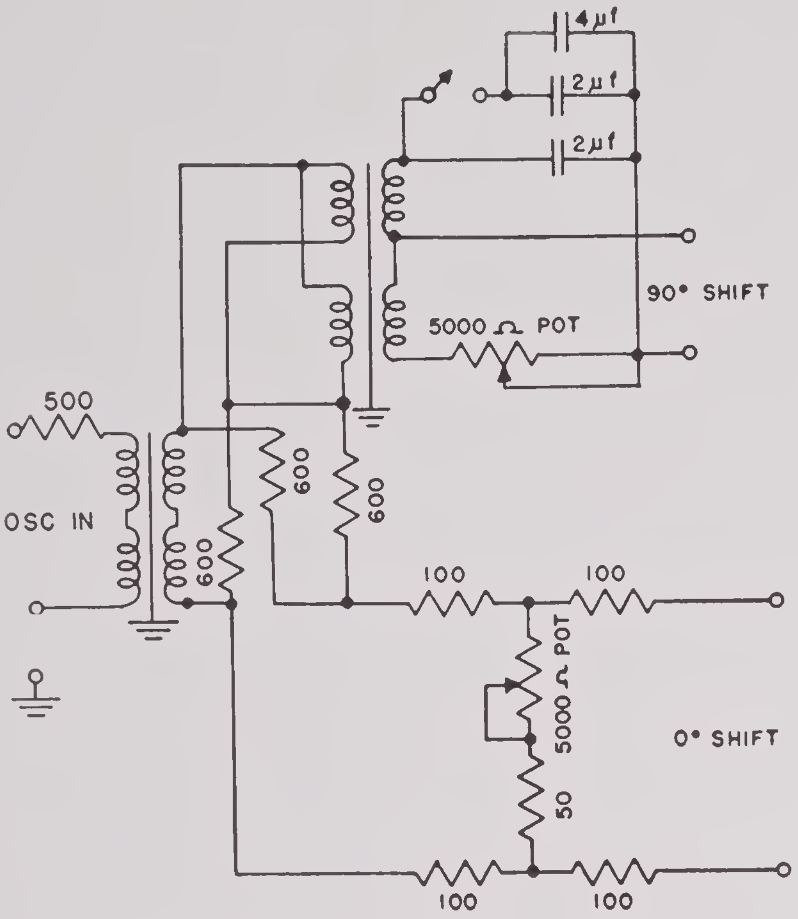

Figure 5. Phase shift circuit for producing two quadrature 60-cycle outputs.

14.3.5 Synchronization Amplifier

The synchronization amplifier is a single-tube amplifier with gain control for adjusting the amplitude of the 60-cycle synchronizing voltage (obtained from a Hewlett-Packard 200D oscillator) that is fed back to the sawtooth generator.

14.4 GENERAL PLAN FOR 0.1- TO 30-MC SCANNING RECEIVER

Following the work described above, plans were completed for a proposed 0.1- to 30-mc scanning receiver, a block diagram of which is shown in Figure 6.[b] The objective was to uncover any fundamental limitation which might be encountered in building a scanning receiver in this frequency range. Accordingly the plan was to minimize chances for spurious signals resulting from modulation products without resorting to unduly high intermediate frequencies and to use existing designs wherever feasible. Designs already completed for the earlier scanning receiver were to be used for the f-m oscillator and all circuits following the second modulator.

The range of the proposed receiver was extended down to 0.1 mc when it was found that this could be done without any major change other than the addition of one input filter and associated beat oscillator. This would provide a universal-type receiver which could be adjusted to monitor any frequency between 0.1 and 30 mc, thereby eliminating the necessity for a separate receiver covering the frequencies between 0.1 and 10 mc.

Experience in laboratory and field demonstrations indicated that a band of 10 mc was probably the largest that could be viewed advantageously at one time. Consequently, the total range covered by this receiver was divided into three bands of about 10 mc each, any one of which could be obtained by selecting the proper input filter and associated fixed frequency oscillator. Circuit design advantages resulting from this division of frequencies are discussed below.

The proposed receiver was a triple heterodyne circuit in which the incoming band of frequencies was converted by means of a fixed crystal oscillator and modulator to an intermediate band of 55 to 65 mc, which in turn was converted in a second modulator to an f-m band having a mean frequency of 40 mc. This conversion was accomplished by means of an f-m wave varying between 95 and 105 mc obtained by tripling the output of an f-m oscillator varying between 31.66 and 35 mc.

The output of the second modulator passed through a narrow-band 40-mc filter to a third modulator where it was combined with a frequency of 40.548 mc. The resulting lower side band was then filtered, amplified and applied to the cathode-ray-tube indicator.

Aural monitoring of radio stations was provided by means of a receiver responding to either a-m or f-m signals and operating from the 55- to 65-mc intermediate frequency, where minimum tuning effort is required to cover all input frequencies. A marking oscillator adjusted to track with the frequency to which the monitoring receiver was tuned was fed into the 55- to 65-mc amplifier circuit as an aid in tuning the monitoring receiver to any station appearing on the cathode-ray tube.

14.4.1 Components for the Receiver

The main components of the proposed receiver are discussed in this section. The progress which had been

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[b] Reports on some of the research which took place during the development of the two receivers described herein have been included in Chapter 12, thus concentrating the present report on apparatus actually designed and built.

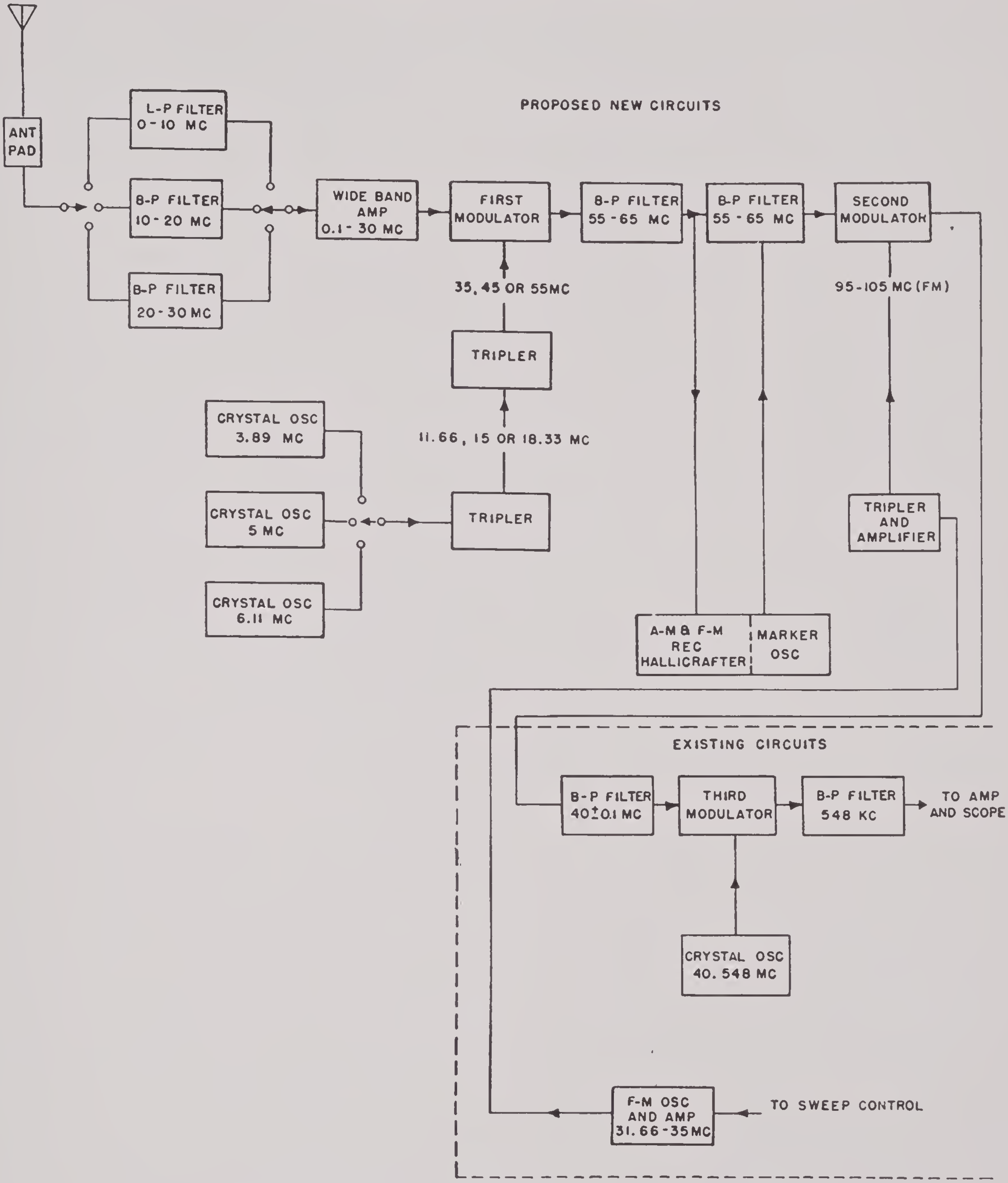

FIGURE 6. Block diagram of the 0.1- to 30-mc scanning receiver.

made on each when it was decided that further work was unnecessary is indicated. Detailed information[c] pertaining to these components not included in this summary, will be found in the contractor's final report.[1]

INPUT FILTERS

Input filters connected between the antenna and the r-f stages limit the range of frequencies connected to the vacuum tubes at any one time to a band width of approximately 10 mc. This not only eliminates image frequency responses but also simplifies the circuit conditions necessary for restricting the modulation products which might produce false images on the screen. Limiting the band scanned to 10 mc further simplifies circuit design and operation by reducing the frequency band swept by the f-m oscillator.

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[c] This information includes insertion-loss characteristics of the input filters, circuit and gain frequency characteristics of the r-f amplifier and first modulator, circuit of the 35-, 45-, and 55-mc oscillator, insertion loss of the 55- to 65-mc band-pass filter, details of interstage transformers and circuit for the 55- to 65-mc band-pass amplifier and second modulator, circuit of the f-m oscillator, circuit of the f-m tripler amplifier, and insertion loss of the 40-mc band-pass filters.

The three input filters were substantially completed.

R-F Amplifier and First Modulator

Experience gained with the 3- to 10-mc scanning receiver indicated a need for greater sensitivity, so an r-f amplifier was provided with a gain of 23.5 ± 1 db over the range of 0.1 to 30 mc. This is only about 7 db more than that in the present scanning receiver, but additional gain was provided in the 55- to 65-mc i-f amplifier so that an overall additional gain of about 25 db was obtained.

The first modulator was arranged for cathode injection of the beat oscillator signal in order to lessen the shunting effect caused by mixing the signals in a pentode. Although the modulator was built with the r-f amplifier, it was tested. The construction is such that conversion to grid injection may be made if necessary with only minor circuit modifications. The output of the modulator goes directly into the 55- to 65-mc filter which constitutes a part of the plate circuit of the modulator.

35-, 45-, and 55-mc Oscillator

It was planned to have the aural monitoring receiver operate from the first i-f frequency, so it was necessary that the beat oscillator frequency remain constant in order that the monitoring receiver might be calibrated in terms of input frequency to the scanning receiver. Consequently crystal-controlled oscillators were chosen to supply the beat frequencies for the first modulator. Crystals for frequencies below 10 mc were more readily obtainable than those for higher frequencies, so crystals of 3.89, 5, and 6.11 mc were chosen. The oscillator output was to be tripled twice, in cascade, to obtain the desired modulating frequencies.

A single oscillator and tripler circuit was contemplated. The frequency desired was to be obtained by switch selection of the proper crystal and associated tuning condensers in the filter circuits. Sharply tuned filter circuits are necessary in order to obtain large discrimination in favor of the desired frequency. This oscillator was not built.

55- to 65-mc Band-Pass Filter

This is a coaxial-line type filter having low loss in the pass band and sharp cutoffs. It had an insertion loss of about 2.5 db over the pass band and was down 10 db at 55 and 67 mc and 30 db at 52 and 70.5 mc.

55- to 65-mc Band-Pass Amplifier and Second Modulator

The input circuit of this amplifier constitutes a part of the preceding filter, so that the filter may operate most effectively. High gain per stage and discrimination in favor of the 55- to 65-mc band, in addition to that obtained by the band-pass filter, was obtained by the use of interstage transformers designed to match the impedances of the associated output and input circuits.

In the second modulator the 10-mc band of frequencies is frequency modulated with 95 to 105 mc. The output of this modulator passes through a 40 ± 0.1-mc filter which is sufficiently narrow to suppress image frequencies in the third modulator. Effectively, the f-m modulating voltage sweeps the input band of frequencies past a filter having a 0.2-mc width. Signals in this band will pass through the filter when the modulation products fall within the limits of the filter.

F-M Oscillator

The f-m oscillator was designed to vary from 31.6 to 35 mc. Its output was tripled to produce the desired modulating frequencies of 95 to 105 mc. This arrangement was chosen in preference to designing an oscillator to work directly at 95 to 105 mc, because it permitted use of the design employed in the earlier 3- to 10-mc scanning receiver. Stability and linearity of operation were enhanced by limiting the f-m sweep to about one-half the maximum obtainable with the circuit used. It was intended that this oscillator should be provided with means for controlling the band width swept and for centering the sweep about any desired frequency in the band, similar to those in use in the 3- to 10-mc receiver.

The f-m oscillator was similar to but smaller and more compact than the one used in the 3- to 10-mc receiver. Preliminary tests showed that its performance was at least as good as that of the earlier model.

F-M Tripler and Amplifier

This is a conventional multiplier-amplifier circuit having the multiplier tube biased beyond cutoff and the multiplier interstage transformers tuned to the third harmonic of the input frequency, which in this case is a band from 95 to 105 mc. One amplifier stage precedes the multiplier stage to insure adequate input level. The f-m tripler and amplifier was built and tested.

Monitoring Receiver and Marker

A receiver providing adequate coverage of the range from 0.1 to 30 mc would need to be divided into five or six separate tuning bands which would necessitate frequent band shifts when operating the scanning receiver. To avoid this frequent shifting, the monitoring receiver was to be operated from the 55- to 65-mc intermediate frequency. The receiver proposed for this use was a special Hallicrafter having an expanded range of 55 to 65 mc and three separate frequency scales of 10 mc each and calibrated in terms of input frequency to the scanning receiver.

A marker oscillator geared to coincide with the frequency to which the receiver is tuned was planned to furnish a signal for injection into the i-f (55- to 65-mc) amplifier as a means for quickly selecting and identifying any signal appearing on the cathode-ray tube indicator.

Miscellaneous

The circuits for the remaining components were to be essentially duplicates of those in the 3- to 10-mc receiver except that the apparatus was to be rearranged to meet the requirements of the new equipment layout. Considerable attention was given to providing proper layout, shielding, and filtering between circuits to eliminate spurious signals such as have been observed in the present model of the scanning receiver.

The 40-mc band-pass filter was built and tested. Two scanning filters were built and tested as part of the study of desirable scanning filter characteristics and for use in the 0.1- to 30-mc receiver. The W-76099 filter consisted of two crystal sections separated by a resistance pad. The W-76132 filter consisted of four crystal sections separated by vacuum tubes or resistance pads.

14.4.2 Further Work

Many problems in connection with panoramic reception remained to be solved but these were beyond the scope of Projects C-27 and C-36. The importance of many of these problems is contingent on the purposes for which the receiver is desired and the operating conditions.

Some of these problems are listed below. No attempt has been made to evaluate their relative importance.

1. Radio-frequency tuning ahead of the first modulator varied in synchronism with the sweep wave to reduce intermodulation products and make scanning-receiver performance more nearly comparable to that of fixed-tuned receivers.
2. Division of the band swept into fractions for reception, amplification, and modulation and then recombination to permit rapid scanning of bands of particular interest without requiring continuous scanning of the entire band.
3. Scanning receivers to indicate primarily relative amplitudes rather than the frequencies of the signals scanned. It seems likely that radically different scanning-filter and indicator characteristics may be desirable.
4. Mechanical versus cathode-ray tube indicators.
5. Facsimile and magnetic-tape recorders.
6. Blanking devices for reducing operator distraction due to the presence of identified signals.
7. Electronically controlled f-m oscillators for use at higher frequencies and to afford larger percentage swing.
8. Investigation of detector characteristics and their effect on receiver resolution, particularly resolution between signals of widely different amplitude.
9. Simplified methods of combining wide- and narrow-band scanning. (Present technique involves separate equipment and indicator.)
10. Electronic switching of broad-band circuits without the present large signal-to-noise ratio penalty.

Chapter 15

RECEIVER FOR PULSE SIGNALS

Research leading to the design and construction of a scanning receiver covering the range of 350 to 750 mc, indicating when radio position-finding pulses are directed at the receiver, arranged to scan automatically, to stop and sound an alarm on reception of a signal, and to indicate azimuth of received signal. Information from this project contributed to better understanding of interception problems and was used in the development of the AN/ARQ-9 radio set.

15.1 INTRODUCTION

Under projects C-27 and C-36 the principles of panoramic reception had been explored quite thoroughly but nothing had been developed for the band width and frequency range contemplated under Project C-39.[a] After a study of the principles involved and the problems, a model receiver was constructed capable of automatically scanning the band of 350 to 750 mc, stopping on reception of a signal, or alternatively continuously scanning the band and sounding an alarm on reception of a signal. Features were included to determine the repetition frequency and pulse length of the incoming signals and arrangements were worked out whereby the azimuth from which the signals came could be determined.

15.2 GENERAL PRINCIPLES

In studying the problem of reception of radar pulses with a scanning receiver, it was necessary to consider the types of information desired and the differences between the scanning reception of radar pulses and that of ordinary signals.

Information which might be portrayed would include existence of pulses, frequency, azimuth, range, pulse rate, pulse shape, and length of pulsing period. These types of information differ both in difficulty of determination and usefulness after being determined. The first four are of greater use tactically; the latter three are of interest more from the research point of view.

The existence of pulses and their carrier frequencies are probably of primary interest, since their very presence indicates the presence of radar equipment. Determination of frequency indicates whether or not the source is friendly. Azimuth indication would be of considerable tactical importance on a ship or aircraft, since it would enable the craft to approach the source of signals.

It was apparent after some study that it would be an enormous undertaking to construct a receiver which would furnish complete information concerning all variables for all types of transmission. Accordingly, it was decided to construct a receiver of fairly simple design which would indicate the presence and frequency of both pulse signals and other types of transmission.

15.3 SCANNING RECEPTION OF PULSE SIGNALS

There are a number of differences between scanning reception of radar pulses and that of ordinary signals. In the first place the frequency spectrum of a h-f pulse is quite broad. For practical purposes most of the energy of the pulsed high frequency may be assumed to lie within a band centered at the carrier frequency and having a width equal to twice the reciprocal of the pulse length. This band for a 1-μsec pulse would be approximately 2 mc.

It is desirable that the scanning filter be at least as wide as the pulse signal band for two reasons. First, this affords a good signal-to-noise ratio and, second, it permits stopping the receiver and viewing the pulse shape. Since the resolution is limited in any case by the signal band width, no sacrifice of resolution results from making the filter band at least as wide as the signal band.

In a panoramic or scanning receiver there is a fundamental relation between the width of the scanning filter and the scanning speed, i.e., the rate at which frequencies are swept past the filter. To obtain satisfactory transient response from the scanning filter, the band width should not be less than that indicated by the following formula:

$$B = K_B\sqrt{nF},$$

where B = width of scanning filter in cycles per second measured between 6 db points, nF = scanning speed in cycles per second per second (i.e., the product of repetition rate n and total frequency band F), and

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[a] Project C-39, Contract No. OEMsr-311, Bell Telephone Laboratories, Inc., Western Electric Co., Inc.

K_B is a proportionality factor. For good design and small level difference between adjacent signals which are to be resolved, the value of K_B may be taken as 1.5.

Although the above relationship constitutes a very important limitation in the reception of narrow-band signals it turns out in a scanning receiver for pulse signals that with practical repetition rates and frequency sweeps, if the width of the scanning filter is made equal to the pulse signal band, the scanning speed is no longer a limitation. Thus, for example, with a 2-mc filter the scanning speed may be of the order of 10^{12} cycles per second per second.

Duty Cycle

A further factor in scanning reception of pulse signals is the low duty cycle, i.e., the fact that the signals are present only a small part of the time. In the process of scanning reception a number of pulses are lost, the fraction received being equal to the ratio of the width of scanning filter to the total frequency band covered. If, in addition to frequency scanning, the receiver employs azimuth scanning or if the source of the pulse signals employs azimuth scanning, the fraction of the original pulses received is much further reduced.

Pulses must be present during a sufficient percentage of the time to actuate the indicator associated with the scanning receiver. The minimum band width for the scanning filter to satisfy this condition may be stated as

$$B = K_T F T,$$

where B is the band width, F is the total frequency band, T is the period of the pulsed signal or the inverse of the pulsing rate, and K_T is the number of pulses that must be received per second to actuate the indicator. If we assume the total band to be swept is 400 mc and that 10 pulses per second are necessary to actuate the indicator, then the scanning filter should be 10 mc wide for a 400-cycle pulsing rate. Furthermore it is apparent that scanning reception in combination with azimuth scanning either in the receiver or in the transmitter presents an extremely difficult problem.

Signal-to-Noise Ratio

The signal-to-noise ratio of a scanning receiver for radar pulses is poorer than that of the same receiver for continuous narrow-band signals of the same total field strength. There are two reasons for this. First, the pulse receiver is responsive to noise all the time, while the pulses are present only a small part of the time. The loss in signal-to-noise ratio on this account corresponds directly to the duty cycle of the radar pulses. Second, the received noise power varies directly with band width so that the noise is correspondingly greater for wide-band pulse power than for a signal of the same power concentrated in a narrow band. Consequently the sensitivity of a receiver for pulses is poorer than that for a receiver for continuous narrow-band signals in direct proportion to their relative band widths.

In addition it was necessary to consider the band width required for stopping the tuning mechanism while the signal was still tuned in. These considerations led to the choice of a scanning-filter band width of approximately 10 mc. The total band covered, 350 to 750 mc, was determined by the sweep range realizable in the beating oscillator. The scanning rate of 1 sweep per second was chosen to provide a suitable interval between audible alarms for different signals expected to be in the range and to allow time for the mechanism to operate. This sweep rate was also satisfactory for later addition to the receiver, if desired, of a visual indicator to show simultaneously all signals present in the frequency band.

15.3.1 Alternative Arrangements Studied

In addition to the design finally selected, other alternative arrangements were studied. The first scheme considered for indicating the existence and frequency of pulse signals portrayed the signals by a vertical deflection of the cathode-ray beam, the height of the deflection being proportional to the frequency of the received pulse. A horizontal time scale would be provided by rotating the tube or by means of a sawtooth sweep circuit. The unfavorable signal-to-noise ratio caused by the fact that the whole band must be amplified and not just that portion occupied by the single pulse being received forced the abandonment of the scheme.

Another arrangement considered would indicate azimuth of the signals as well as their existence and frequency. Analysis indicated that very few signals would be received because of the fact that both receiving and transmitting antennas would be rotating, that the receiving antenna would have high directivity, that the receiver band width would be a small part of the whole frequency band swept through. This scheme was abandoned.

15.3.2 Azimuth Indicating System

The model receiver actually constructed was designed so that it could be used as the basis of a four-channel azimuth scheme in which four antennas with circular directivity patterns directed to the four compass points fed voltages to the four deflecting plates of the cathode-ray tube. The radial deflection of the beam indicated the direction from which the signals came.

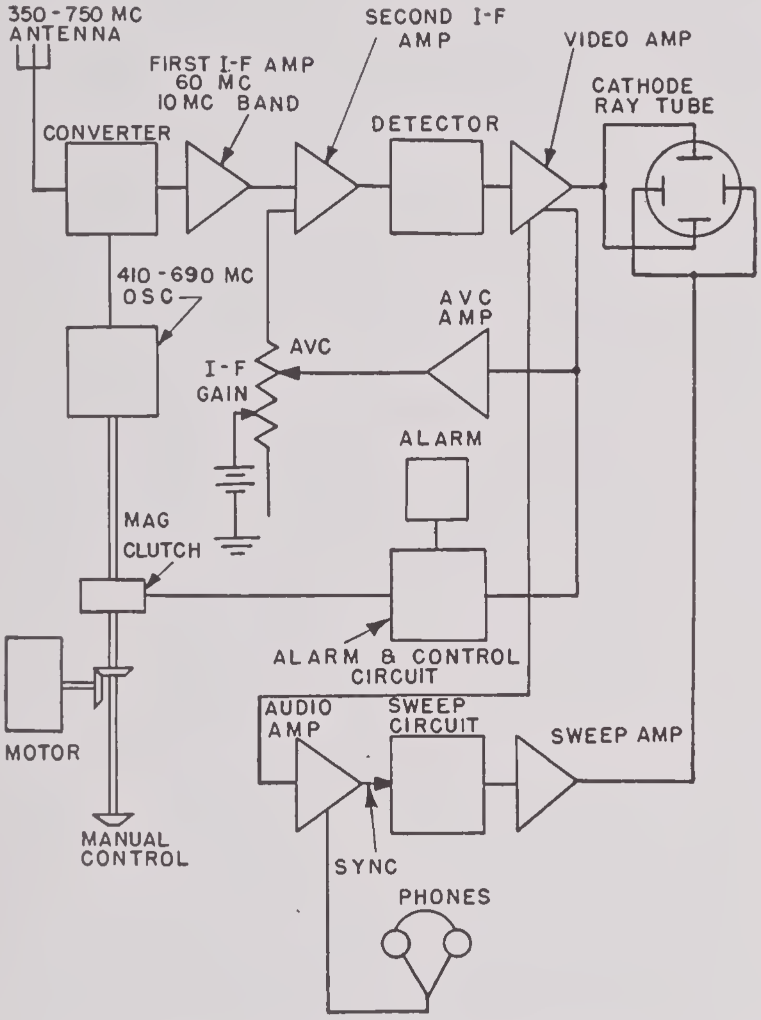

Figure 1. Block schematic of the 350- to 750-mc panoramic receiver for pulse signals.

15.4 DETAILS OF MODEL RECEIVER

A block schematic diagram of the receiver is shown in Figure 1. The heterodyne oscillator is controlled either by the motor and clutch arrangement shown or manually. In addition to the automatic volume control, a manual gain control has been provided to adjust the threshold of sensitivity of the receiver. Part of the signal from the video amplifier is connected to the alarm and control circuit which actuates a bell when a signal is received and is also capable of stopping the sweep oscillator when a signal is received. Part of the output signal is fed to an a-f monitoring amplifier. Part of the audio frequency from this amplifier is connected to a linear sweep circuit to synchronize the sweep rate with the pulsing rate of an incoming signal.

Antenna

The antenna provided with the receiver is mounted in the middle of the top of the cabinet and consists of a quarter-wave vertical stub resonant at approximately the middle of the band, 550 mc. The antenna was made with a length-to-thickness ratio of about 6 to provide reasonable performance over the 350- to 750-mc band of the receiver.

Converter

The converter circuit is shown schematically in Figure 2. The converter itself is a Western Electric fixed silicon crystal. A transmission-line stub transformer is connected between the antenna jack and the crystal to match the impedance of the antenna approximately to that of the crystal. The transformer was designed for an impedance ratio of 72 to 275 ohms. This transformer also serves as a broadly tuned input filter to attenuate the signals outside the 350- to 750-mc band. A 20-$\mu\mu$f capacitance is part of the mounting block of the crystal converter. This capaci-

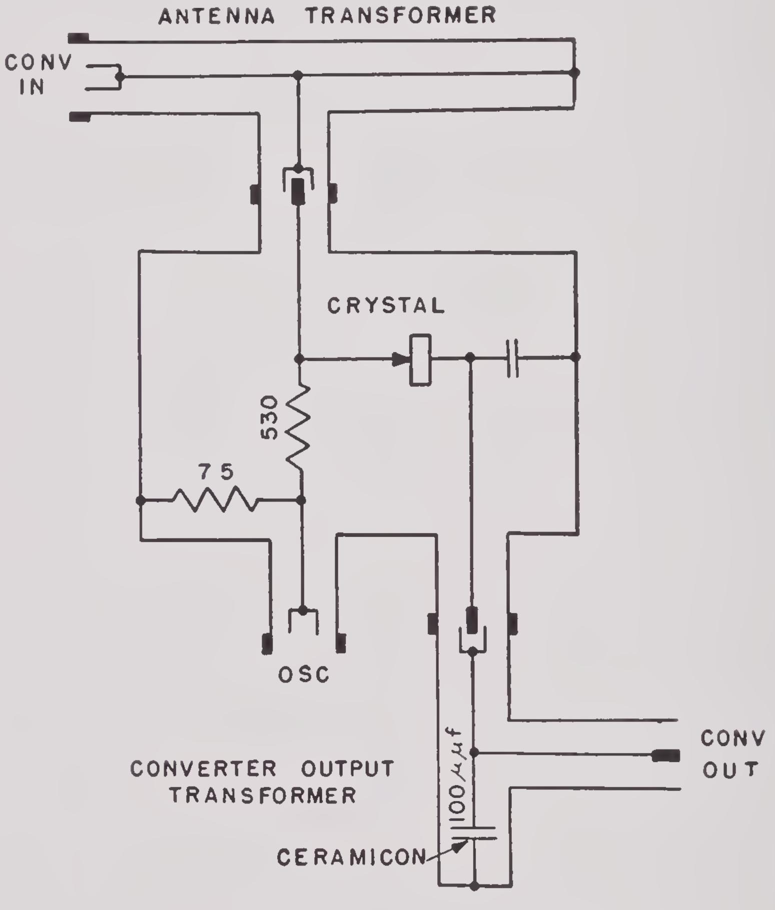

Figure 2. Silicon converter schematic.

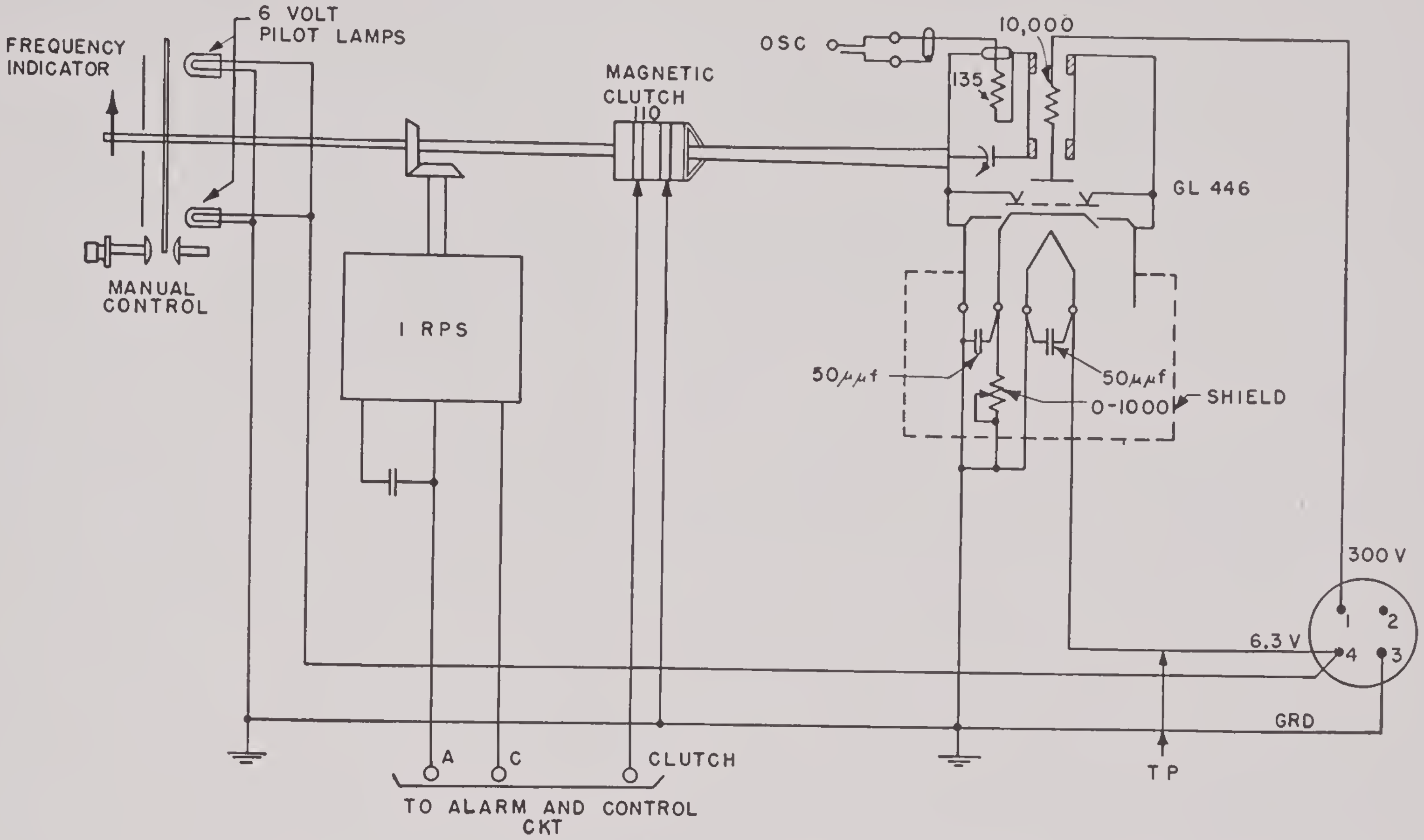

FIGURE 3. Oscillator using GL-446 tube tuning over range from 420 to 690 mc.

tor is effectively a by-pass for the signal frequencies. The 60-mc i-f output from the converter is obtained across this capacitor. Another stub transformer, this time at 60 mc, is used to match the 275-ohm impedance of the crystal to the 72-ohm impedance of the first i-f amplifier.

OSCILLATOR

A schematic of the oscillator is shown in Figure 3 and a cross section in Figure 4. The GL-446 tube is mounted in one end of a cylindrical cavity which is effectively a short-circuited coaxial line about λ/4 long

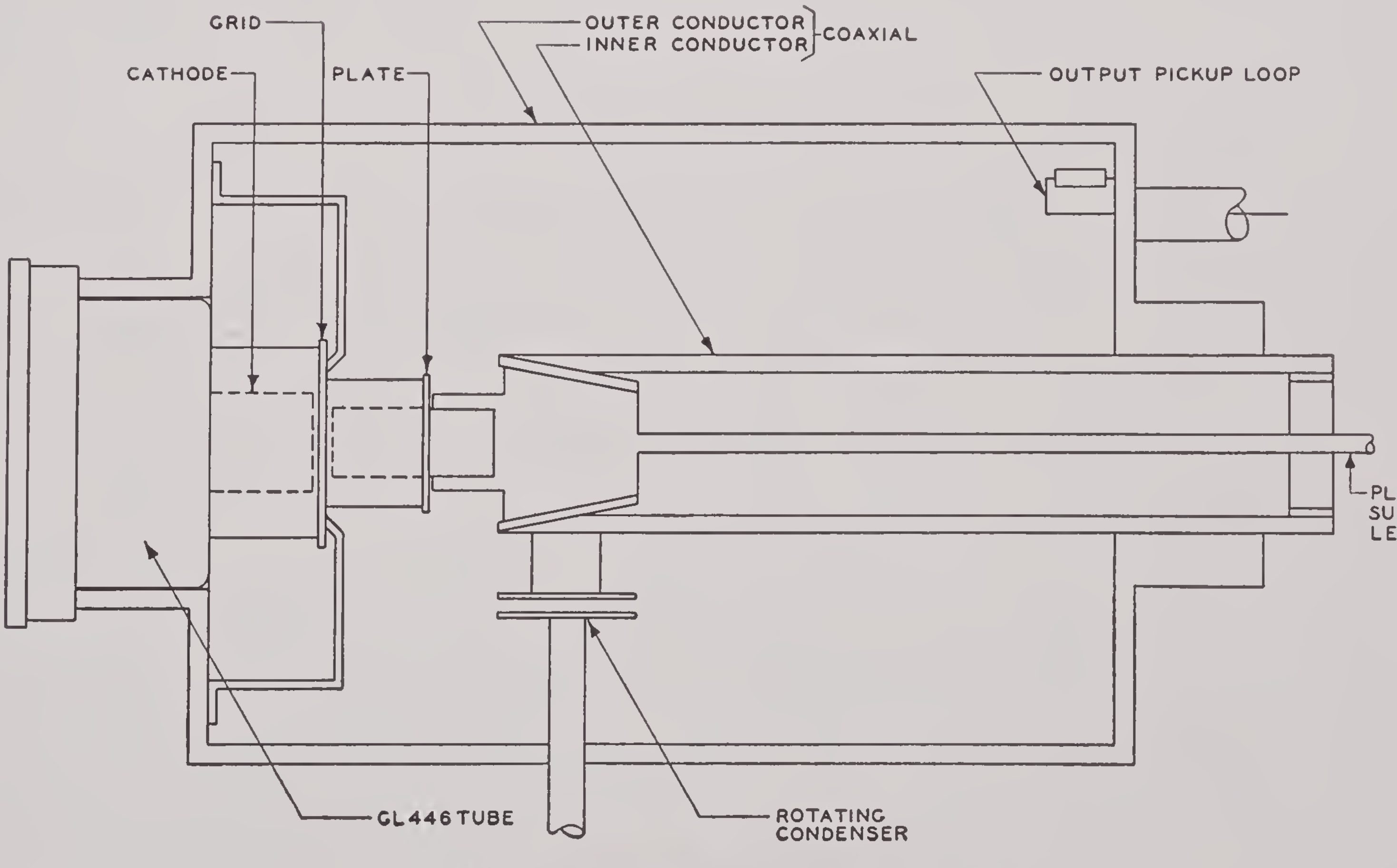

FIGURE 4. Cross section of the oscillator tube and cavity.

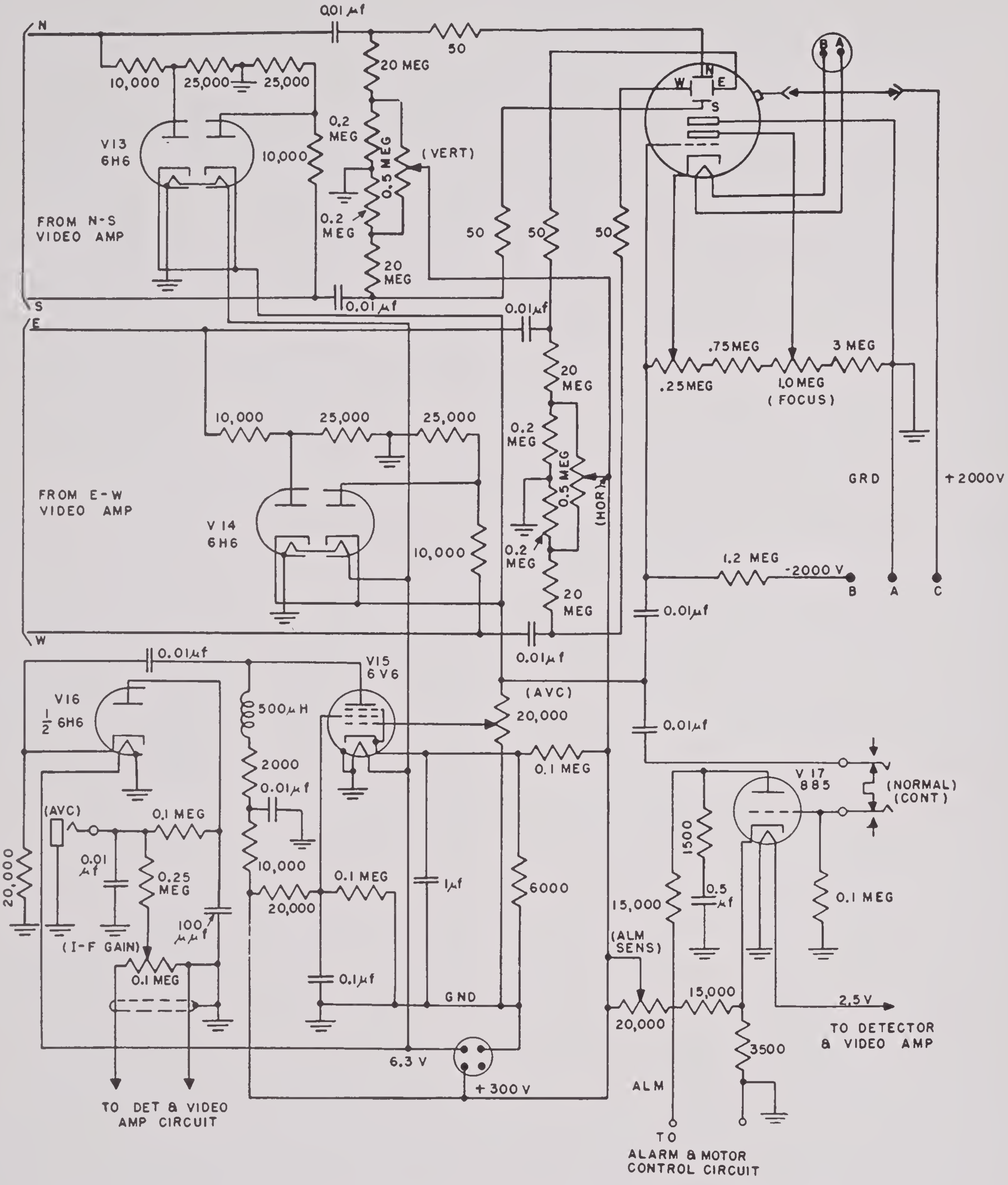

Figure 5. CRO control circuits and gas tube circuit for alarm and motor control.

at the highest frequency of oscillation, 690 mc. This coaxial line is connected between the plate terminal and cathode of the tube. Grid excitation for maintaining oscillations is obtained from two thin metallic fingers which project from the end of the cavity and make contact with the grid connection on the tube. These fingers effectively form pickup loops in the oscillating field at the high-voltage end of the cavity.

The oscillator is tuned over a frequency range from 690 to 410 mc by a variable air capacitor connected across the cavity at its high impedance point, that is, as near to the tube as possible. The capacitor has two semicircular plates and in 180 degrees of rotation its capacitance varies about 5 $\mu\mu$f. This capacitor may be operated manually or driven by a motor.

The plate voltage for the oscillator is supplied through the inside of the coaxial inner conductor. The plug which fills the end of this hollow inner conductor has a capacitance to the inner conductor of about 100 $\mu\mu$f acting as a high-frequency by-pass from the plate to the cavity. Additional filtering is obtained by virtue of the inductance of the plate lead wire and another by-pass capacitor near the other end of the inner conductor. The output is obtained by means of a small pickup loop projecting into the cavity at the high-current or short-circuited end.

Intermediate-Frequency Amplifiers

The two i-f amplifiers are much the same in construction. The first amplifier has two stages, the second four, 713A tubes being used throughout. Both amplifiers have 72-ohm input and output impedances. The first amplifier has a gain of about 30 db and the second a maximum gain of 65 db. The gain of the second amplifier is controlled by means of the grid bias voltage on the second and third stages. This voltage is furnished by the a-v-c circuit or the manual gain control which is a potentiometer across a 9-volt battery. The combined gain of the two amplifiers may be varied between 88 db and substantially 0 gain by a range of 9 volts grid bias.

Detector

The detector is of the so-called infinite impedance type using a 713A tube. The envelope voltage appears across a 5,000-ohm resistance in the cathode circuit of the tube. The input impedance to the circuit is 72 ohms, similar to that of the input to the i-f amplifiers. A choke in the output lead is provided to block the 60-mc intermediate frequency.

Video Amplifier

The envelope of the received signal is applied to the input of the video amplifier which is a 6AC7 tube. The proportion of the signal which is connected is controlled by means of the video gain controls in the grid circuit of the 6AC7 tubes. The cathode bias resistor for the two 6AC7 tubes is not by-passed and provides phase inversion for the applied signal. The plate-circuit stages for the two 6V6 output tubes are inductance-compensated resistance networks to extend the gain characteristic of the amplifier.

The plate voltages and cathode biases of the tubes in the video amplifier are all connected to a rotary switch through suitable dividing resistances so that a voltage indication may be obtained on a 0- to 2-ma meter on the face of the panel.

A-V-C Circuit

Output signal pulses of positive polarity from the video amplifier are connected through the two halves of a 6H6 tube V13 and applied to the grid of the 6V6 tube V15 as shown in Figure 5. The proportion of the voltage applied to the 6V6 tube is controlled by means of the a-v-c potentiometer. The amplified impulse is rectified by V16 and applied to the grids of the second and third stages of i-f amplifier No. 2. The amplifier gain is thus controlled by the amplitude of the received signal. The i-f gain potentiometer controls the fixed bias and hence the maximum gain of these stages.

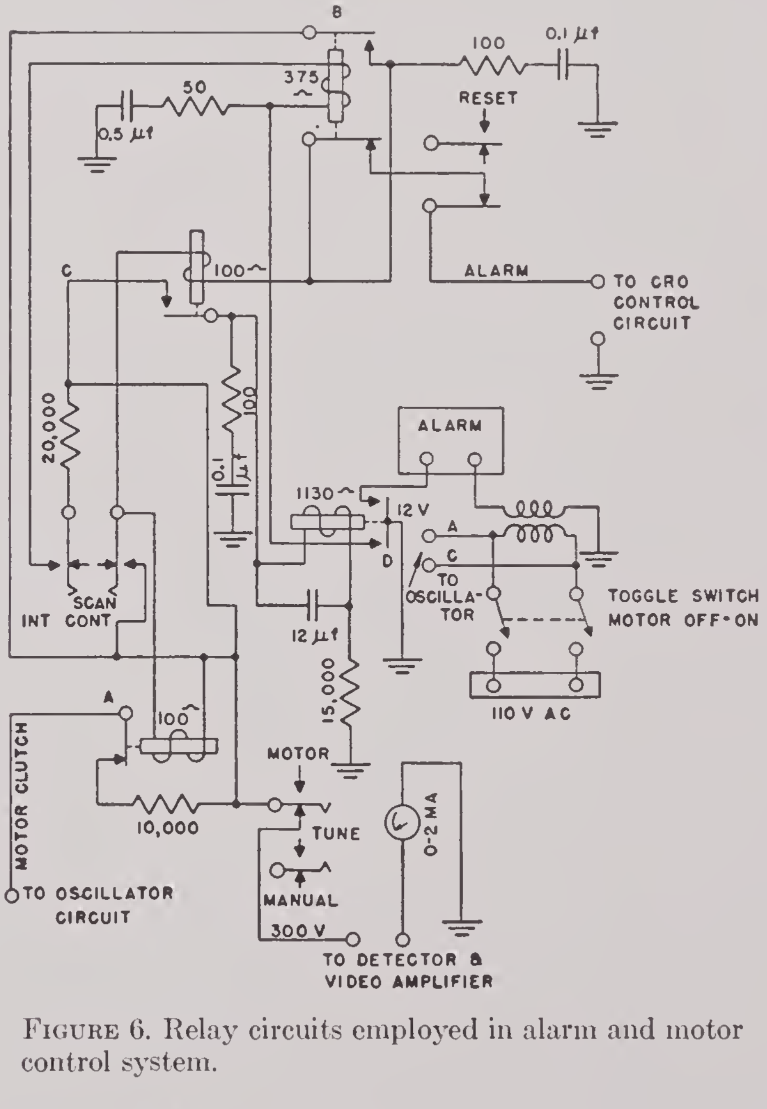

Figure 6. Relay circuits employed in alarm and motor control system.

Alarm and Motor Control

A gas discharge tube and a relay sequence permit continuous scanning of the frequency range, sounding an alarm whenever a signal is swept through or, to permit interrupted scanning, sounding an alarm and automatically stopping the scanning mechanism on the receipt of a signal.

Positive pulses from either side of the output of the video amplifier are connected through a 6H6 and applied to the grid of the 885 gas discharge tube V17 as shown in Figure 5. The breakdown point of this tube is controlled by the "alm sens" potentiometer. Figure 6 shows the alarm and motor control circuits.

With the scan key in the "cont" position for continuous operation, the discharging of the gas discharge tube by a signal causes the *C* relay to operate. This, in turn, operates the *D* relay, which completes the alarm circuit. The *D* relay also operates the *B* relay, restoring the gas discharge tube and releasing the relays by opening the plate battery lead. Thus, the alarm has sounded and the trigger circuit restored to normal for a subsequent signal response.

With the scan key in the "int" position for interrupted operation, the receipt of a signal causes the gas tube to discharge, the *C* and *D* relays to operate, and the alarm to sound as described for continuous operation. In this case, however, the *C* relay also operates the *A* relay, which releases a clutch mechanism disengaging the motor from the frequency scanning capacitor of the oscillator. The total sequence of operation is sufficiently rapid to permit the scanning oscillator to remain tuned to the incoming signal. To release the circuit to resume the scanning for other responses, the reset button is pressed, opening the plate circuit of the gas tube.

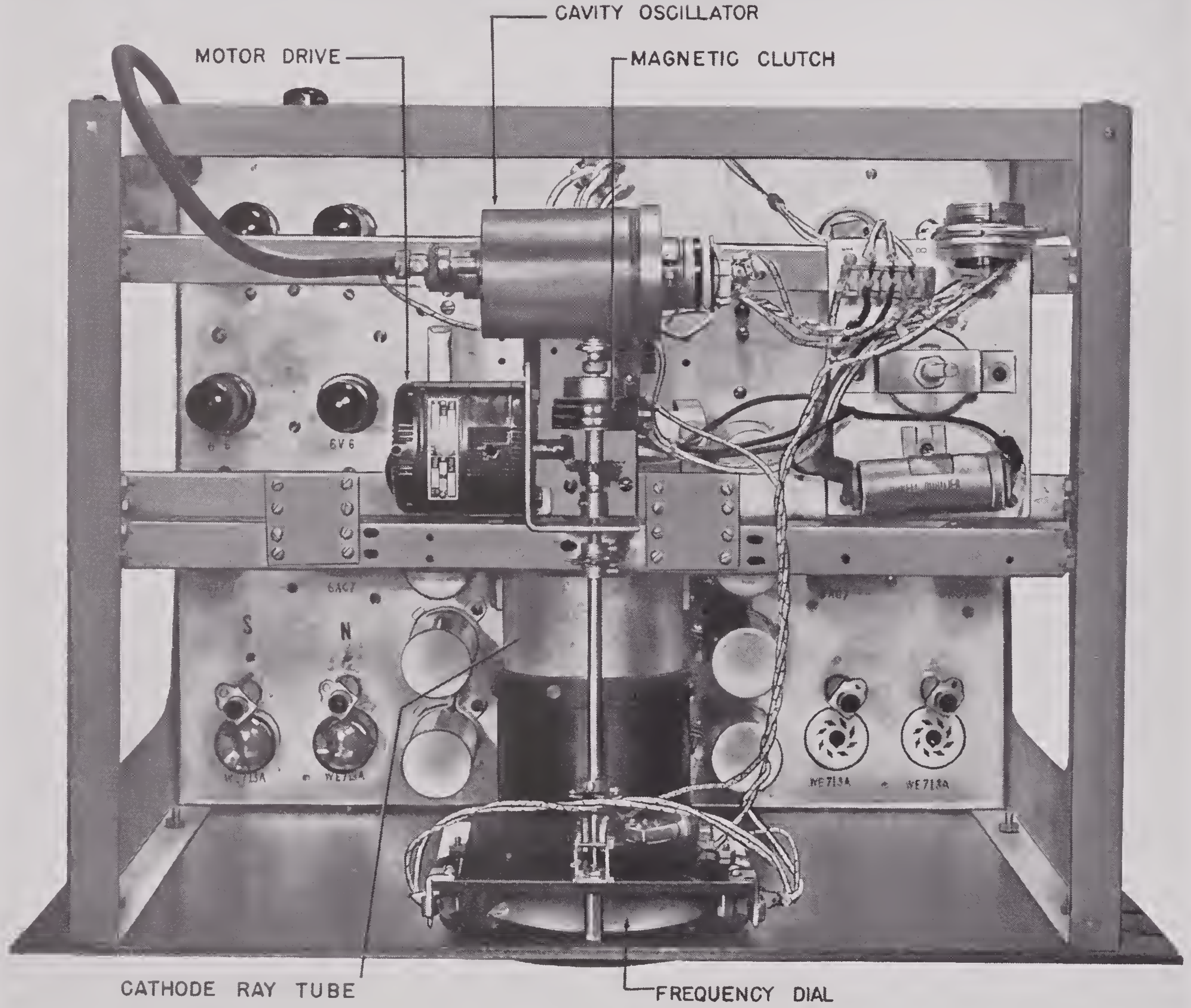

FIGURE 7. Oscillator and video chassis of the 350- to 750-mc pulse receiver.

Operating the "alm" key to the off position opens the grid lead of the gas tube disabling the alarm and motor and control circuit. When operating the receiver manually, the tune key is operated to the "man" position. This disengages the motor clutch mechanism and disables the alarm and relay circuit. Operating the manual frequency control also opens the circuit to the clutch mechanism.

Monitor and Sweep

A monitor and sweep circuit are provided to assist in the identification of signals and the determination of the type of transmission. Connection to this circuit is made from the cathode of the video tube V3. The signal is amplified by a 6AC7 tube and then is applied to the grid of a gas tube sweep circuit, the output of which is applied to the horizontal deflecting plates.

Test Oscillator

Pulse-modulated signals are supplied from a Wien or RC oscillator for testing the receiver. Sine-wave frequencies of 400, 1,000, 1,600, 2,000, and 4,000 cycles from the oscillator are selected by a switch and are amplitude-controlled by a lamp in the cathode circuit of the oscillator. The bridge output signals are peak-chopped and are then applied to parallel resonant circuits tuned to 1, 0.5, 0.33, and 0.2 mc respectively. A diode damps out all but the first positive half cycle of the transient so that the width of the resulting pulses is 0.5, 1.0, 1.5, or 2.5 μsec. These pulses are amplified and shaped and then modulate the plate voltage of a cavity oscillator, permitting it to oscillate only during the pulse interval. The frequency range of this oscillator, which is similar to the oscillator in the receiver, is 550 to 730 mc.

The oscillator build-up time is just less than 0.5 μsec so that the pulses resulting are that much shorter than the impressed modulating pulse. Operating the pulse-cont key to the "cont" position removes the modulating signal, permitting the cavity oscillator to supply a continuous output. In this condition the oscillator will cover the range of 450 to 730 mc. A quarter-wave stub antenna, similar to that used with the receiver, may be connected for transmitting.

Receiver Characteristics

The sensitivity of the receiver to pulse-modulated signals is such that about 50 μv are required for a ½-in. deflection. The band width of the i-f amplifier is 9.5 mc at the —3 db point, frequencies 10 mc away from the 60-mc mid-point being 70 db down. At a pulsing rate of 4,000 cycles and a pulse width of 2 μsec, a 56-db variation in input signal produces a 16-db variation in cathode-ray tube deflection.

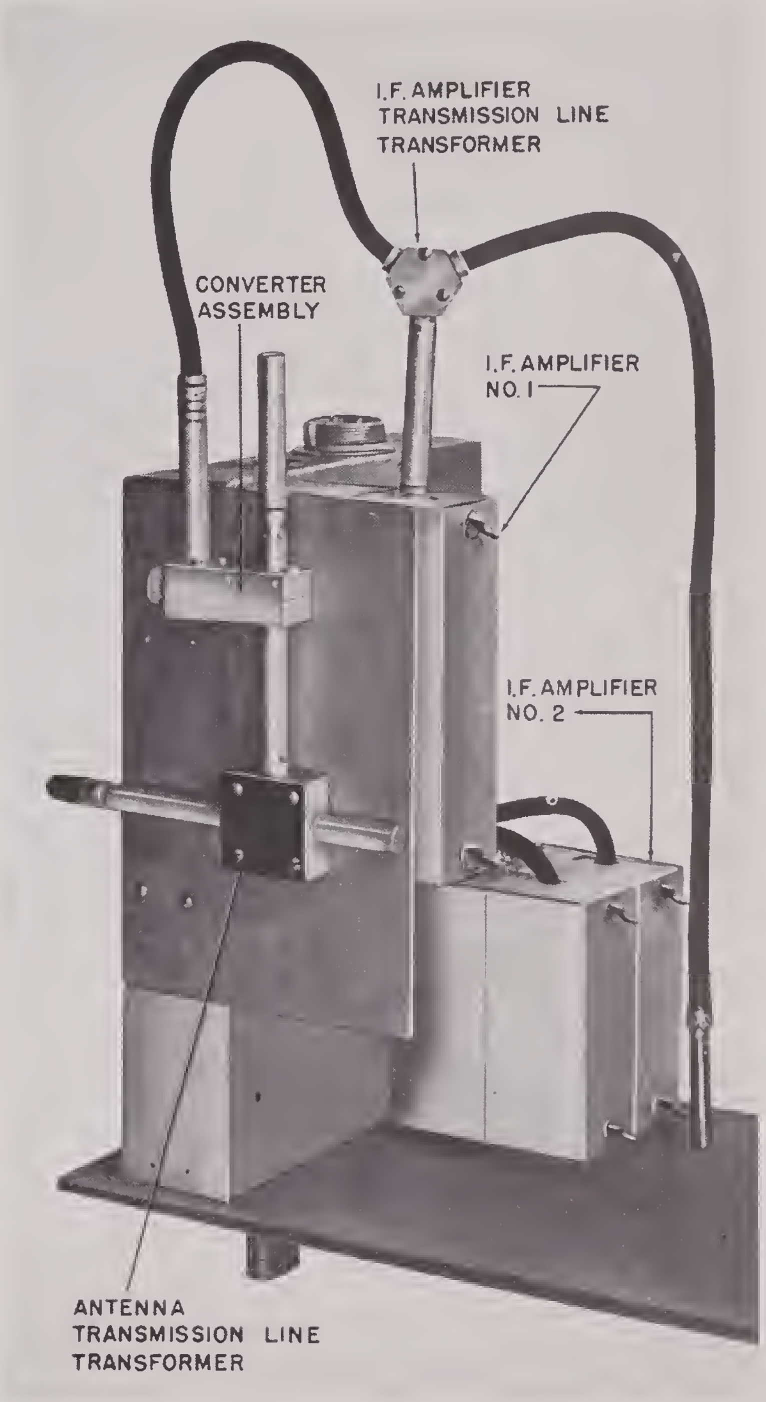
I.F. AMPLIFIER TRANSMISSION LINE TRANSFORMER
CONVERTER ASSEMBLY
I.F. AMPLIFIER NO. 1
I.F. AMPLIFIER NO. 2
ANTENNA TRANSMISSION LINE TRANSFORMER

Figure 8. Converter assembly for panoramic receiver for pulse signals.

15.5 FUTURE WORK

There are a number of possibilities for further work along lines related to this project.[1] Some of these represent improvements which might be made on the equipment as it now stands and others represent extension of the work in different directions.

A panoramic indicator might be attached to the present receiver to give an instantaneous cross section of the frequency band covered. This would be in the usual form of a vertical deflection for each received signal on a horizontal base line to which a frequency scale could be applied. The circuits necessary to do this have been designed. The scan would involve a 360-degree potentiometer which would be driven by the oscillator motor drive mechanism. From this potentiometer a horizontal pyramidal sweep wave would be generated. This would be applied to the horizontal deflecting system of the cathode-ray tube instead of the sawtooth sweep which is used for examining signals. The received signals would be applied to the vertical deflecting plates as at present. Because of the image response, the frequency scale for the abscissa would be exactly the same as the frequency scales on the upper half of the frequency dial of the receiver. That is, there would be a frequency scale for signals higher in frequency than the beating oscillator and another scale for signals lower in frequency than the beating oscillator. These two scales differ from each other by 120 mc or twice the intermediate frequency. This double frequency response makes the panoramic indication of doubtful value in the model receiver.

The equipment layout and arrangements of this receiver were designed with a view towards the possibility of constructing a 4-channel receiver for azimuth indication.

PART V

INTERFERENCE GENERATION

The extreme importance of proper communication between units and between commanders and their units in any form of warfare, is paralleled by the equivalent importance of detecting enemy communications and, if possible, of making them ineffective. Thus, observers of World War II before our entry into it were struck by the effectiveness of the liaison between tanks and planes by which German tanks were piloted by German planes flying over the battlefields and were warned of hazards and obstacles in their path.

If this team of plane and tank could be broken up by making it difficult or impossible for one to communicate with the other, the value of the team could be decreased considerably.

Several studies of this problem were made in Division 13 and resulted in at least two jammers in addition to a fundamental examination of the most effective means of disrupting communication by speech or code on amplitude or frequency modulation by facsimile or pulse transmission.

Three Division 13 projects were concerned directly with the interference production problem, C-25, C-26, and C-56. The first two were devoted essentially to production of interference generators covering the bands from 2 to 20 and from 15 to 30 mc respectively, although C-26 also covers a great deal of basic research, while C-56 dealt entirely with the fundamental aspects of the means for effective jamming. In the summaries that follow the work is presented in a logical sequence of interest rather than in the chronological order in which it was done.

Chapter 16

STUDY OF INTERFERENCE GENERATION

Research to determine the most effective types of interference signals for jamming radio-telephone and telegraph communication channels. This summary contains all of the technical data available in the contractor's final report.[1] Later, more extensive work was done by Division 15.

16.1 INTRODUCTION

This summary presents in condensed form the data and conclusions from the work on this project,[a] including those from some preliminary experiments which preceded the actual authorization of the project. The experimental work covered rather completely the question of the effectiveness of various a-f noises in rendering speech and telegraph signals unintelligible.

16.2 SPEECH TRANSMISSION

Four general types of a-f interference were tested:[2] resistance noise, speech, noise of the scanning type whose fundamental frequency is modulated so as to sweep through the voice band in various manners, and noise consisting of trains of impulses. In addition, the noise generated in the generator developed in Project C-26 was studied. The data were obtained in terms of the per cent of discrete sentences rendered unintelligible by the noise for various noise-to-signal ratios expressed in decibels, and were supplemented by spectrograms of typical conditions, showing visual evidence of the degree of masking. The data from the intelligibility tests are summarized in Table 1.

The general conclusion from the work is that *for a noise to have maximum effectiveness its spectrum must be continuous both in time and in frequency over the band.* Any appreciable gaps in either dimension permit unmasked fragments of speech to be heard which convey considerable intelligence. A supplementary conclusion is that visual inspection of its spectrogram will usually indicate the general effectiveness of a given noise. At least it can be stated positively that if the noise spectrogram is of an open nature so that the pattern of superposed speech would be readily visible through it, the noise will be relatively ineffective in suppressing speech intelligence.

The effective noises of those tested were resistance noise and mixtures of speech from two to twelve voices. These were about equally effective and required that the interference exceed the desired speech signal by about 10 db, as read by a volume indicator, for complete suppression of intelligence under the quiet listening conditions of the experiments. The curves were quite steep. If the noise was lowered 10 db below the above maximum, about nine out of ten sentences could be understood. Impulse and scanning types of noise were all much less effective, requiring from 7 to 29 db higher noise levels than the above for equal loss of intelligibility.

TABLE 1. Noise-to-signal ratios necessary to jam speech.

Type of noise	N/S ratios* in db for sentence errors of		Peak factor† (db)
	50%	90%	
Resistance noise	4	8.5	10
Speech interference			
Babble of 12 voices	4.5	9	
Mixture of 2 voices	2	7	
Scanning types of noises			
Stepped tones	19–27		
Sawtooth scanning‡	16	(23)	
Impulses			
Sharp impulses			
51 per sec	33		19
118 per sec	17	25	15
Buzzers			
180 per sec	11	16	10
325 per sec	15	(25)	9
350 per sec	15		6
Resistance noise interrupted at rate of 12 per sec§			
0% of period	4	8.5	10
25%	11	(17)	11
50%	30	36	13
75%	33	37	16
90%	28	32	20

*N/S ratio refers to the difference in db between the rms noise power referred to one milliwatt and the speech volume read on a standard volume indicator in the standard manner. Figures in parentheses were derived by considerable extrapolation beyond the experimental data.

†Peak factor is the height of the instantaneous peaks in db above the rms value of the wave. For comparison, the peak factor of speech is about 13 db and of a sine wave is 3 db. If peak rather than rms values of the noise are of interest, the given N/S ratios should be increased by the peak factor.

‡Consists of a tone rich in harmonics the fundamental of which is periodically swept nearly linearly from 2,000 to 400 cycles with instant return. Figures hold for scanning rates of 5 to 45 cycles per second.

§The interruption was not quite complete but consisted of the sudden introduction of a loss of 30 to 38 db. The background noise, rather than the pulses, was controlling for interruptions greater than 50 per cent of the period.

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[a] Project C-56, Contract OEMsr-626, Bell Telephone Laboratories, Inc., Western Electric Co., Inc.

If the noise spectrum is continuous, the loss of intelligibility is due to a true masking phenomenon dependent only on the hearing mechanism of the ear. That is, the listeners actually fail to hear the speech sounds at certain rather critical superposed noise levels. This can be explained by the accepted theory of hearing, for such a noise continuously stimulates all of the nerve endings on the basilar membrane in the cochlea and therefore interferes with the transmission to the brain over any nerve path of impulses recognizable as speech. The threshold of understandability of speech in the presence of this noise is therefore about the same for all normal persons. On the other hand, if the noise is discontinuous or does not possess energy at all frequencies, some of the nerves at least some of the time are unstimulated by the noise and are free to transmit speech signals to the brain. The energy level of such noise must therefore be considerably greater to interfere with intelligibility, and moreover the listeners are aware that they can still hear the speech sounds even when they no longer can understand them. It was found that observers differed greatly in their ability to understand speech through such noise, since it depended largely on their psychological ability to disregard loud noise and to listen to and interpret the weak speech sounds that, being unmasked, can be heard through the noise. Experience may greatly improve a listener's ability to hear through this kind of noise.

In the tests involving the use of noise consisting of speech from two voices, an attempt was made to add a psychological handicap to the observers by using for one of the voices the same voice (from a recording) that later spoke the sentences for the intelligibility tests and, moreover, repeating a similar list of sentences. In spite of this, the speech interference was not noticeably more effective than resistance noise, in terms of the peak powers required as measured by a volume indicator which integrates over about 0.3 second. A single voice was less effective than resistance noise because of the gaps between syllables, words, and sentences. The two- and twelve-voice interferences were equally effective within about 2 db and this doubtless would apply to any other intermediate number of voices.

A further comment on impulsive noises is warranted. If a noise with a continuous spectrum, such as resistance noise, is interrupted for small percentages of the time, it rapidly loses in effectiveness. For example, Table 2 shows the ratio by which the average power of periodically interrupted resistance noise must be increased over that of uninterrupted noise for the same degree of jamming. The second column indicates the required power averaged over the interruptions and the last column the corresponding power in the pulses or, in other words, before interruption.

TABLE 2. Comparison of interrupted and uninterrupted resistance noise.

% of total time Noise is interrupted*	Required factor by which power must be multiplied Power averaged over long time	 Power during pulse
0	1	1
12.5	2.25	2.6
25	5	7
50	400	800
75	1,000	4,000
90	1,000	10,000

*The rate of interruption was four times per second for the 12.5% case, and 12 times per second for the others.

Two important conclusions are apparent.

1. If it is necessary to interrupt the interference in order to observe the wave to be jammed, the interruptions should be for as small a per cent of the total time as practicable.

2. Nothing can be gained by causing a given band of interference to sweep across several channels so as to hit any one only part of the time (assuming that time constants of limiters, a-v-c circuits, etc., in the receivers do not affect the results). For example, if the noise is swept across four channels, it is on any one only 25 per cent of the time and is interrupted on each 75 per cent of the time. The figures for 75 per cent interruption in Table 2 show that if the noise is distributed in this way among four channels, the total power must be multiplied by over 4,000 as compared with steady noise on one channel. If, instead, it were continuously applied to all four bands, the required total power would be increased only proportionally to the band covered, or by a factor of only 4. The latter method is evidently the more efficient one by a ratio of more than 1,000 to 1.

16.3 TELEGRAPH TRANSMISSION

Two types of interference were tested for their effectiveness in preventing experienced operators from reading hand-sent telegraph signals of a pitch of about 935 cycles.[3,4] One interference consisted of resistance noise and the other was a sine wave whose frequency suddenly changed at intervals of 0.1 second among nine values lying between 580 and 1,500 cycles, one of which differed from the telegraph frequency by only 5 cycles. In the case of resistance noise, reception was completely interfered with when the noise intensity or power per cycle was 25 db below the marking telegraph signals, regardless of the band width, provided it was greater than 120 cycles. If the band were limited to only 120 cycles, complete loss of intelligibility could be obtained with a total noise power which is about 4 db weaker than the marking telegraph signal. The stepped frequency, when listened to through a 3,000-cycle band, completely interfered with the signals when its power was about 1 db stronger than the marking signals. When listened to through a 120-cycle band, which removed some of the interfering frequencies, the interference power had to be increased about 10 db for equal effectiveness.

Telegraph signals were plainly visible in spectrographs through resistance noise when the noise power in a 3,000-cycle band exceeded the marking telegraph signal power by 5 and by 6.5 db, at which values the errors were 20 and 60 per cent, respectively. At 10 db, however, where the errors were 100 per cent, the signal is invisible. In the case of stepped frequency interference, the signals were still visible at a level where they could not be read. This indicates that failure in this case was due to the mental confusion created by the interference rather than to a masking phenomenon.

Of the two interferences tested, the resistance noise was appreciably superior in effectiveness when the band was narrow. It, of course, has the advantage that if it can be assured that the noise falls somewhere on the telegraph signals, it is immaterial what part of the noise band is close to the signal. The masking is caused by the noise within 50 cycles or so of the telegraph frequency and this has the same character at all parts of the spectrum. It is possible that another telegraph signal located within a very few cycles of the signal to be jammed would be more effective than either of the noises tested. The practical possibility of achieving this depends upon other characteristics of the system, however, and it was felt that further tests should be made on a radio rather than an audio basis.

16.4 RESISTANCE NOISE

If resistance noise should be adopted as the interference signal, the question of its generation and transmission would become important. One method which seems promising is to generate it at voice frequency (which is readily done by means of a gas-tube circuit) and to apply this to modulate an f-m transmitter. It is obvious that this would be a very effective interference against f-m channels, since it would have a maximum effect on the limiters in the receivers and since the noise heard in the receiver outputs would be pure resistance noise. To obtain an idea of its probable effectiveness against a-m channels, the radio-frequency spectrum of the interference signal was determined theoretically.[5] It was found that the relative energy per cycle versus frequency is substantially a normal distribution curve centered about the unmodulated carrier frequency. This curve would have a standard deviation corresponding to the frequency deviation which would be caused by impressing on the transmitter a steady (d-c) current of the same value as the rms value of the impressed noise. It appears that this noise would therefore be quite effective against a-m if the standard deviation is adjusted to correspond to the channel band width.

From the practical standpoint, this method of generating the interference would be very flexible, as the band width can be readily widened or narrowed to fit particular channels to be jammed by merely changing the gain in the a-f circuits of the interference transmitter, thus changing the degree of modulation.

Chapter 17

RADIO INTERFERENCE GENERATORS

Design and development of interference generators[a] (jammers) covering the 2- to 20-mc and 15- to 30-mc bands. This summary covers work performed on Project C-25[1] prior to May 20, 1942, and on Project C-26[2] prior to June 29, 1942, after which respective dates both projects were transferred to Division 15 where more extensive work was done on this problem.

17.1 FUNDAMENTAL STUDIES

In the book *Speech and Hearing*[3] the masking of pure tones by other pure tones, as well as the masking of the former by complex sounds, is discussed. As is generally the case in this type of subject, no simple all-inclusive rules can be given. Certain generalizations, however, can be made.

1. A low-frequency tone can obliterate a high-frequency tone if the former is at a sufficient level (say 100-db sensation level).
2. A high-frequency tone in general cannot mask a low-frequency tone.
3. The threshold shift for a masked tone (required increase in level above original threshold value in order that it may again be perceived) is greatest when the masking tone is close to it in frequency. In this special case, it is possible for a higher-pitched tone to mask one of lower pitch, a contradiction or exception to 2 above.

The reason advanced for the greater effectiveness for masking of a low-frequency tone is that if it is of sufficient intensity it is able to produce subjective overtones in the ear due to the nonlinearity of the latter, and one or more of these overtones will be close to the high-frequency tones to be masked, and hence (by 3) very effective in doing so. The reason advanced for 3 is that the masking tone has stimulated the particular nerve fibers of those terminating in the basilar membrane that normally respond to the frequency to be masked, and caused them to discharge their unit loads, so that they can no longer respond to the frequency to be masked.

4. A complex sound produces further masking by virtue of the cross-modulation effects (summation and difference beat frequencies) as well as by means of the harmonic overtones which it creates subjectively in the ear.

No conclusions are presented in the above reference as to the relative effectiveness for masking of single tones and complex sounds. However, the loudness of a sound whose energy is distributed over the spectrum appears greater than that of a sound of equal intensity whose energy is concentrated in a narrow band, provided the hearing mechanism is thereby subjected to sound levels of 40 db or higher.[4] This might indicate that a complex tone is more efficacious for masking than a single tone where such masking is desired for a range in the spectrum rather than for one particular frequency.

17.1.1 Masking by Clicks and Pulses

Isolated and recurring clicks and pulses appear to gain more rapidly in loudness with increase in level than a steady 1,000-cycle tone.[4] This would indicate that for effective masking, the interfering sound should be relatively high in intensity. More quantitative data will be presented below. Measurements of the effect of disturbing noises and single-frequency tones upon a telephone headset which has a fairly flat frequency response indicate that for pure tones, the 1,000-cycle tone has the most disturbing effect, although for the telephone set employed frequencies from 330 to 3,500 cycles per second are within 10 db of the 1,000-cycle tone in disturbing effect.

17.1.2 Masking by Noise

Noise in the frequency range containing the transmitted speech most important to understanding interferes with understanding more than noise in some other frequency range, although it is true that certain noises produce a harmful effect through annoyance rather than through masking.[5]

Experimental tests[3] indicate that the threshold shift of speech sounds produced by noise is about the same as the average shifts produced on a 500-, 1,000-, and 2,000-cycle tone. In another paper,[6] it is stated that in the region of 1,000 cycles a tone of one frequency can just be perceived in the presence of thermal noise when its intensity is about 60 times the noise intensity per cycle for noise bands no narrower than 60 cycles. Thus, for example, for a 6-kc band, the signal need be only about 1/100 of the total noise intensity, i.e., 20 db below the latter.

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[a] Project C-25, Contract OEMsr-89, Farnsworth Television & Radio Corp.; and Project C-26, Contract OEMsr-285; Federal Telephone & Radio Corp.

17.1.3 Masking of Code Signals

An unpublished memorandum[7] throws considerable light on the interference of telegraph code signals. This report indicates that for a 3,000-cycle band width, thermal noise must be 10 db stronger than the signal to cause a sharp break in the ability to receive the message. Complete obliteration of the signal required that the noise be 17 db stronger than the signal. An alternative form of disturbance employed was that of nine tones, 580, 1,010, 1,320, 810, 1,100, 940, 1,260, 650, and 1,500 cycles, produced in sequence at a rate of approximately one such group per second. This "jumping" tone for the 3,000-cycle band width, had to be but 2 db higher than the signal to disrupt the transmission. The above frequencies require only about one-half the band width that the thermal noise was permitted to occupy.

The telegraph pitch was 935 cycles. According to 3, most effective masking would be accomplished by frequency components close to this in value. This seems to be borne out by a test in which a 120-cycle band-pass filter (935 cycles mid-frequency) was used in cascade with the rest of the system. Actually this filter was employed to simulate the sharp tuning capabilities of a receiver, but the results indicated that the narrower 120-cycle band of thermal noise was as effective as the original 3,000-cycle band in masking the signals when it was 4 db *below* the signal in intensity. This indicates, as stated above, that only the components close to 935 cycles are effective in masking, and that those remote from this frequency represent waste energy as far as interference is concerned.

17.1.4 Further Discussion

The band-pass filter mentioned above was also employed with the jumping tones. The input to the filter in this case had to be increased from the original ±2 db to ±11 db. This is probably due to the fact that only the 940-cycle tone came through with any appreciable intensity, although some of the other tones could be heard. While some interesting conclusions might be assumed from this report, it must be remembered that it is only a preliminary one and is subject to further verification. For instance, if only the 940-cycle tone came through the 120-cycle filter with any appreciable intensity, then the effective interfering energy was essentially in pulses of 1/9-second duration, coming at a repetition rate of one per second. The average energy would be 1/9 of the peak. The increase in required energy when the filter is employed is 9 db, which is eight times the energy. This is close to nine times the energy required during 1/9 of a second to give the same average energy as when the filter was not employed. Thus it might be argued that the increased level for the jumping tone was due simply to the fact that the band-pass filter cut down the energy transmitted. On the other hand, many of the tones were remote from the telegraph signal in frequency and hence probably not very effective in masking the latter. We may be permitted to conclude, however, that the only virtue of noise over any other type of interference is that no matter how limited the receiving channel is made in frequency range, there will be available noise components in that range, since noise is a random signal whose spectrum consists of components of all frequencies and all of equal magnitude.

17.1.5 Criteria for Interference

All the references cited, except the last, discuss the complete obliteration or masking of tones. For successful interference it is not necessary to mask completely, but only to reduce the intelligibility of the transmission below an acceptable limit. In commercial practice, this limit is 60 per cent successful transmission. Probably a lower limit is acceptable in military practice, but the exact value does not seem to be known yet. The interference level for this purpose will be less than that for complete obliteration, how much less depends upon whether the test is one for disconnected words (discrete word intelligibility) or for disconnected sentence (discrete sentence intelligibility).[8] If complete masking is obtained, it is of course obvious that no intelligence can be transmitted.

In connection with these facts it should be noted that the interference curves are quite steep. For example, the range from 50 per cent to substantially 0 per cent successful transmission is covered by a change in the level of the interfering resistance noise of only 7 or 8 db for speech and about half that for telegraph.

The exact value of the limit chosen as acceptable is therefore not very important practically speaking. Whether one chooses 60 per cent or 40 per cent successful transmission as the limit affects the required interference by only about 1 db, which is small compared to the other uncertainties involved.

17.1.6 Work Done under Project C-26

One conclusion which had been drawn early was that the form of modulation of the interference generator was of secondary importance. At first pulses of essentially rectangular form and varying width were employed, interrupted periodically to permit monitoring of the enemy signal. This form of modulation was subsequently changed to one in which the pulses varied in pitch at a low rate, say one every 2 seconds. The width of the pulses was adjustable, as was the rate of "warbling."

Articulation tests indicated that discrete word intelligibility was only about 30 per cent for the latter form of modulation as compared to about 50 per cent for the former type, which was an appreciable, but not outstanding, improvement. The interfering sound in these tests was about 6 db above the speech sounds. Tests at Fort Monmouth gave similar results after the tuning adjustment was improved.

The final form of modulation described above occupies the region of speech frequencies, is of complex form and sufficiently irregular to be considerably distracting and hence annoying. It thus essentially meets the requirements for interference detailed above. Moreover, it is produced by a simple circuit employing two thyratron tubes, and, as stated previously, can be adjusted as to pitch and as to rate of warbling. Further development along this line did not appear justified at the time.

17.1.7 Military Considerations

Of greater importance than the foregoing are the military considerations involved. Among these are the questions as to whether several communications are to be jammed simultaneously or only one at a time, whether the link consists of two parties or more than two, whether the frequencies involved are fixed (crystal-controlled transmitters) or continuously adjustable, and also what they are, whether switching from one frequency to another is possible or not, and whether a-m, f-m, c-w, or i-c-w signals are to be jammed.

17.1.8 Requirements Met by Project C-26 Interference Generator

Since the project was fundamentally exploratory in nature, equipment was built to meet the following requirements: monitor a band of frequencies, about 3.0 mc wide at one time, cover a range of 15 to 30 mc in two bands, jam a-m, f-m, c-w, or i-c-w signals, and handle communication systems in which switching from one frequency to another is employed by the enemy to avoid interference.

17.1.9 Electrical Considerations

The discussion on the masking of tones dealt with the effects of interfering sounds upon a signal after these had been conveyed to the ears of the observer. There are, however, electric circuits to be traversed before the composite of signal and noise reaches the ears, and it is in this portion of the system that a great amount of ingenuity has been exercised in the past to obviate interference. The precautionary measures employed must now be circumvented.

17.1.10 Tuning Requirements

For interference to be effective, the jammer must be accurately tuned to the enemy frequency. Tests made in the laboratory and confirmed by those performed at Fort Monmouth indicated that even in the case of a-m speech reception, accurate frequency alignment was necessary, since otherwise very little interference energy would get into the system unless the impinging energy was prohibitively great. The alignment had to be within 1 kc or better.

17.1.11 Original Method of Tuning

The original method of frequency alignment was that of having a spectrum-scanning receiver, which portrayed during one scan the frequency to be jammed (among many others that might be on the air) and during the next scan the frequency of the interference generator. Alignment was deemed adequate when the two signals (resonance curves) were placed in juxtaposition. To facilitate this, a narrower scan was employed for a final adjustment.

17.1.12 Present Method of Tuning

This, however, was found to be inferior to a method subsequently developed. The wide scanning was retained and gave the operator a visual indication of the stations on the air and told him whether or not the enemy signal to be jammed had shifted to a new frequency to elude interference. For fine tuning, the horizontal sweep on the oscilloscope screen was retained and serves as a linear time base during those periodic intervals of time when scanning occurs. The receiver, however, is "fix-tuned" during this time to the enemy signal by a finer adjustment of the tuning capacitor. The result, during fine tuning, is that the enemy carrier appears periodically on the screen, and the nature of the signal, whether a-m, speech, or c-w, can be easily determined.

Apparently, the reason why the scanning method is not sufficiently accurate is that the interference signal is in the form of pulses unrelated in timing to the small interval when the scanning capacitor is tuning through the interference carrier frequency. As a result, sometimes two pulses, sometimes three, occur during such an interval, and hence the resonance peak changes in amplitude in an erratic, jumpy manner, which makes the alignment of it with the center cross hair a difficult matter. On the other hand, a much larger number of pulses appear during one entire scan (1/50 second) and hence the picture in the tuning position is steadier.

17.1.13 Appearance of Signals on Screen

The transmitter portion of the interference generator is ganged with the receiver with a vernier control to permit it to be more accurately tuned to the receiver, which has already been fix-tuned to the enemy signal. The interference signal also appears on the oscilloscope screen in alternate scans to those presenting the enemy signal. All this is possible because the transmitter is keyed on and off periodically. During the time it is on, it swamps out the enemy signal in the receiver. (The receiver is blocked by a negative voltage during this time, so that it will not be overloaded by the transmitter.) During the time the latter is off, the enemy signal has a chance to come through, especially since the sensitivity of the receiver has been restored by the removal of the blocking voltage.

The two signals appear superimposed upon the screen. Complete alignment is attained when both signals appear at maximum amplitude. The method is simple and accurate but unfortunately gives no indication as to the direction in which the enemy may have changed frequency to elude interference nor his new position in the spectrum. Hence the spectrum scanning is normally used to monitor the enemy signal, and fine tuning used momentarily after his signal has been located by the former method.

17.1.14 Interference for Frequency Modulation

In the case of frequency modulation, a further reduction in interference is possible if the desired signal is at least 6 db higher than the interference. On the other hand, if the latter is 6 db higher than the former, it will take control of the limiter and tend to obliterate the signal. Tests performed with a fixed carrier of fixed amplitude seemed to corroborate this, but if the carrier pulsated in amplitude as described above, the control was insufficient, particularly on the Army f-m transceivers.

The transceivers have a 10-μsec release time on the limiters. If the interfering carrier is amplitude-modulated in that it is keyed on and off, then the short receiver release time will permit another signal, such as that of the enemy, to take control of the receiver during the intervals when the interfering carrier is off. The enemy signal can thus come through with, at most, a flutter, but nevertheless intelligible.

By wobbling the interference carrier frequency during the on periods, jamming was practically complete. Apparently the carrier, when on, not only obliterated the signal, but produced a tone as well, which was sufficiently distracting to mask the enemy signal during the times that the latter was on and the tone was off. In this way, amplitude modulation could be jammed without wobbling the frequency and frequency modulation merely by turning on, in addition, a motor-driven rotating capacitor, thus wobbling the frequency. The number of operations to be performed could thus be kept at a minimum.

It was undesirable to wobble the frequency for a-m interference, since the energy was spread out over an unnecessarily wide portion of the spectrum, and hence less was available at the frequency of the a-m carrier to be jammed. For a frequency deviation of about 16 kc, the reduction in power at the carrier or average frequency was found to be about 15 db.

17.1.15 Effect of Noise Limiter

A preliminary investigation pointed to the desirability of operating the transmitter power oscillator so as to obtain momentary peak powers greatly exceeding the normal output of the tube. This could be accomplished by periodically keying the tube with pulses. As first constructed, the tube was keyed for about 1/50 second and off for 3/50 second, during which time scanning and monitoring of the enemy could be performed.

This, however, proved to be an incorrect attack on the problem because of the electrical characteristics of receivers designed to circumvent such disturbances (ordinarily static pulses), that is, noise- or crash-limiter circuits. These may be of the form of a diode, biased by the average value of the carrier so that it is inoperative until the carrier exceeds twice its average value. Static pulses, which are momentary pulses of high amplitude, will operate the diode, which shorts them out. Such a circuit will also short out high interference pulses of the type described above without affecting the gain of the receiver during the off periods of the transmitter. If these latter periods are each of 3/50-second duration, sufficient enemy signal can come through during these times to nullify the masking tone during the other 1/50 second, since it has been greatly reduced in amplitude by the noise-limiter circuit. Such was found to be the case in tests performed at Fort Monmouth.

If no noise-limiter circuit is employed but the AVC has a sufficiently fast release time, then the receiver may recover its gain during the 3/50-second period to permit a sizable enemy signal to come through too.

In any event, the solution is to reduce the peak power for a given average power and permissable plate dissipation by prolonging the on period for the transmitter and reducing the modulation voltage. This was done: the on period was made 3/50 second, and the off period 1/50 second, which is sufficient time for monitoring. The interference was much more effective after this change was made.

17.1.16 Degree of Modulation

Finally, it is evident that maximum interference sound output is produced in the receiver if the latter is operated at maximum effect, that is, if the interference carrier is modulated 100 per cent in the case of amplitude modulation or over the maximum deviation capabilities of the receiver in the case of frequency modulation. (More than 100 per cent amplitude modulation will be removed by the noise limiter.) Such modulation characteristics are now more nearly met by the 3/50 second on and 1/50 second off characteristic.

17.2 GENERAL DESCRIPTION OF THE C-26 JAMMER

The interference generator developed under Project C-26 consists essentially of a tunable transmitter and scanning receiver powered by an inverter fed from a 24-volt battery, with transmitter and receiver ganged together mechanically and synchronized in time so as to afford an interval for jamming and one for monitoring the enemy signal.

The inverter drives a scanning capacitor and a switch in the r-f tuner so as to generate synchronizing, blanking, and sawtooth waves in addition to tuning the receiver cyclically over approximately a 3-mc range in the spectrum.

The sawtooth wave furnishes suitable horizontal deflection for an oscilloscope. The blanking signal blanks out the return trace on the oscilloscope screen. The synchronizing signal locks a multivibrator in the modulator unit at half its frequency for keying and modulating the transmitter unit.

The output of the modulator actuates the transmitter so that it is on for one period of time and off for a shorter period, during which the enemy signal may be observed. A portion of the transmitter output is picked up by the receiver to afford observation of the interference signal on the oscilloscope screen as well as the enemy signal, but in alternate sequence. To prevent overloading of the receiver when the interference signal is on, the gain of the receiver is reduced at such times by a blocking voltage, also obtained from the modulator unit.

17.2.1 Results Secured

Field tests indicated that for reasonable distances between jammer and enemy receiver an interference power of about 25 watts is required for jamming the usual field transmitters with an output of about 5 watts. This power is confined to a band width not exceeding 5 kc and much less than this for CW. Thus the power required for jamming a single channel is not very great.

If, however, a wider band is to be jammed, for example, 0.5 mc, simultaneously, then the power must be 2,500 watts; for a 1-mc range it must be 5 kw; and if the enemy shifted frequency as much as 5 mc the power required to disrupt his communications in spite of his frequency shifts would have to be 25 kw, which would be too great for mobile operation.

If the energy is spread by employing narrow modulation peaks (pulses) then a noise limiter in the receiver will partially or perhaps completely vitiate the effectiveness of the jammer. The use of a spark transmitter with its well-known wide band was suggested[2] as an interesting line of development that might yield profitable results.

The final report on Project C-26[2] contains other discussion of such matters as narrow-band interference, the advantages of separation of keying and modulation, of the use of a common antenna, of the virtues of using two receivers, one for scanning and one for tuning, of using motors to eliminate mechanical coupling of receiver and transmitter and other pertinent factors leading to the design of effective jammers.

17.3 WORK DONE UNDER PROJECT C-25

The purposes of Project C-25 were as follows.

1. To investigate the possibility of locating an enemy transmission accurately and quickly within a band approximately 20 mc wide.
2. To jam the enemy within a relatively narrow band (3 kc) without interfering with other communications.
3. To investigate the type of transmission which would most effectively jam i-c-w communications.
4. To accomplish this within a range of a few miles with minimum power.
5. To construct complete airborne equipment in the minimum space and weight required.

17.3.1 Accomplishments

A panoramic receiver was used to spot the enemy. This receiver was suitable over the 18-mc range and three bands. Various widths of spectrum could be examined by expanding the sweep. Spotting was by visual indication but phones were provided for listening to determine the advisability of jamming.

Having spotted the enemy, a transmitter was tuned to the required frequency and was energized by a trigger. The transmitter signal appeared on the panoramic receiver opposite that of the enemy so that the transmitter could be tuned accurately before pressing the trigger.

A "step c-w" type of modulation was employed in which the transmitter frequency was shifted within 4 kc at a random rate. It was found, however, that a random tone, modulated and applied to the step transmitter signal, was more effective.

Normal communication signals could be jammed at a distance of approximately 10 miles from a plane 5,000 ft in the air.

The complete equipment weighed approximately 100 lb and occupied about 5 cu ft.

17.3.2 The End Results

The effectiveness of jammers of the types developed under Projects C-25 and C-26 may be made evident by the following excerpts from reports on their use during field maneuvers in Louisiana. "A complete mechanized division of the Army was delayed two hours in starting operations because its communications were jammed by the equipment. It was necessary to resort to motorcycles for communication before the operation could get under way." From another communication, "... the presence of the equipment was not known to anybody except the highest ranking officers. It proved to be brilliantly successful and disrupted maneuvers so completely that the men in tanks had to stop operations and instead got out of their vehicles and ate their lunch!"

After these tests in this country, some 50 of the sets, covering various frequency ranges, were ordered for use at once on the battlefields, even though the demonstrations indicated that improvements of considerable value could be made.

PART VI

RADIO TRANSMISSION FORECASTING

CONFIDENTIAL

Chapter 18

IONOSPHERE STUDIES

Five laboratories participated in a cooperative project on radio transmission conditions, measuring the field intensities of numerous transmitting stations on various frequencies from 660 to 15,355 kc, and continuing measurements on conditions in the ionosphere. At the completion of individual contracts, work was continued under the Interservice Radio Propagation Laboratory [IRPL], set up by the Joint U. S. Communications Board to furnish comprehensive radio propagation service to the Armed Services.

The following summary is written from the final reports of the individual projects, the bibliography showing the relation between the several contracts and their extensions. Related work will be found in Part II of Volume 1, Division 13, on direction finding and antennas.

18.1 INTRODUCTION

At the time the projects[a] were started there had been no systematic attempt to forecast the radio and ionospheric storms or disturbances which interrupt long-distance h-f radio communication. Sufficient background of experience had been acquired, especially at National Bureau of Standards [NBS] and at the Department of Terrestrial Magnetism, Carnegie Institution of Washington [CIW] to indicate the promising nature of a broad study of the type described below. Some success had already been achieved in short-time forecasting of radio disturbances, using very limited data.

Two general types of data were collected under the several projects, one dealing with geomagnetic phenomena (magnetic activity) and solar phenomena and the other with ionosphere characteristics and field-intensity measurements. The solar data were collected by a chain of solar observatories which submitted to the Department of Terrestrial Magnetism daily reports which were correlated with available magnetic information (Project C-53). For the ionosphere and radio data, NBS, under Projects C2-13 and C2-49, acted as the centralizing agency.

18.2 ACCOMPLISHMENTS

A brief summary of the work accomplished under the several projects will be found below but the entire effort may be summarized as follows: Improvements were made in methods of determining maximum usable frequencies, lowest usable frequencies, skip distances, and distance ranges. A method of calculation of the absorption index over a path was devised. A new service of short-time forecasting of radio transmission conditions was developed, providing forecasts for a week ahead. World charts showing predicted variation of maximum usable frequencies based on observations at all of the cooperating stations were prepared.

The methods and practices developed were turned over for use of the IRPL, set up by the Joint U. S. Communications Board to furnish comprehensive radio propagation service to the Armed Forces. At the end of the fiscal years of the projects, the work was fully under way and NDRC was able to close out its work relating to radio propagation predictions as a completed phase of the research, sufficient to permit the beginning of an operational program.

Cooperating Laboratories

The laboratories participating in the program of the ionosphere and radio field-intensity observations were National Bureau of Standards (centralizing laboratory), Washington, D. C.; Carnegie Institution of Washington, College (near Fairbanks), Alaska; Louisiana State University, Baton Rouge, Louisiana; Stanford University, Palo Alto, California; University of Puerto Rico, San Juan, Puerto Rico.

These stations observed field strengths and ionosphere conditions of the following transmitting stations:

Station	Frequency in kc	Location
	Recorded at Baton Rouge, La.	
WWV	5,000	Beltsville, Md.
W8XAL	6,080	Mason, O.
COCH	9,435	Havana, Cuba
XEWW	9,500	Mexico City, Mex.
WWV	15,000	Beltsville, Md.

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[a] C-9, no contract, National Bureau of Standards [NBS]; C-13, no contract, NBS; C-14, Contract OEMsr-200, Carnegie Institution of Washington [CIW]; C-20, Contract OEMsr-227, Stanford University [SU]; C-22, Contract OEMsr-378, Louisiana State University [LSU]; C-44, no contract, NBS; C-45, Contract OEMsr-632, University of Puerto Rico; C-46, Contract OEMsr-573, LSU; C-47, Contract OEMsr-590, SU; C-48, Contract OEMsr-558, CIW; C-49, no contract, NBS; C-53, Contract OEMsr-594, CIW.

Recorded at College, Alaska

KGEI	7,250	San Francisco, Calif.
KEP	9,480	San Francisco, Calif.
KGEI	9,550	San Francisco, Calif.
WLWO	9,590	Mason, O.
JAP	9,595	Tokyo, Japan
DJW	9,650	Zeesen, Ger.
WRCA	9,670	Bound Brook, N. J.
JAP	9,672	Tokyo, Japan
DJX	9,675	Zeesen, Ger.
WNBI	9,690	New York, N. Y.
WRUW	9,700	Boston, Mass.
GSD	11,750	Daventry, Eng.
WRUL	11,790	Boston, Mass.
JVZK	11,815	Tokyo, Japan
WBOS	11,870	Boston, Mass.
GSI	15,260	London, Eng.
KWID	15,290	San Francisco, Calif.
KGEI	15,330	San Francisco, Calif.
WGEA	15,330	Schenectady, N. Y.
KWU	15,355	San Francisco, Calif.

Recorded at San Juan, Puerto Rico

WJZ	770	New York, N. Y.
WWV	5,000	Beltsville, Md.
COCH	9,435	Havana, Cuba
GLH	13,525	Dorchester, Eng.

Recorded at Palto Alto, Calif.

WWV	5,000	Beltsville, Md.
CBRX	6,160	Vancouver, B. C.
COCH	9,435	Havana, Cuba
XEWW	9,500	Mexico City, Mex.
GSD	11,750	London, Eng.
WRUL	11,790	Boston, Mass.
JZJ	11,800	Tokyo, Japan

Recorded at Washington, D. C.

WEAF	660	New York, N. Y.
WLW	700	Mason, O.
WBBM	780	Glenview, Ill.
WCKY	1,530	Covington, Ky.
WQXR	1,560	New York, N. Y.
WWV	5,000	Beltsville, Md.
W8XAL	6,080	Mason, O.

Data secured from such radio field-intensity measurements furnish a means of checking, from actual operating conditions, the computations of maximum usable frequencies and are useful in confirming the existence of ionosphere storms and in indicating their intensities and the areas and frequencies affected. They are the chief source of detailed information about sudden ionosphere disturbances and indicate the time of their beginning, their intensity, and their duration. They can be correlated with solar and magnetic data to provide additional information for the prediction of radio transmission conditions. They are used to determine the amount of absorption of sky-wave transmission for different latitudes, frequencies, and lengths of paths, so that the decrease in field intensity in addition to the inverse distance factor may be estimated.[b]

The final report on Project C-19[11] gives the mathematical background for calculating additional attenuation of radio waves by absorption, for determining maximum usable frequencies (see also the final report, Project C-9, *Radio Transmission Handbook*),[1] lowest usable high frequencies and distance ranges. Certain of this analysis is over-simplified and many techniques have been greatly improved and knowledge expanded since the dates of these projects. This analysis, however, gives a clear, concise, and readable picture of the ionosphere and of the manner in which, at the time of writing, ionosphere and field-intensity data could be usefully applied to radio communication predictions.

18.2.1 Apparatus Employed

In general, the equipment used to obtain field-intensity records consisted of a communications receiver in conjunction with a continuous automatic recorder. The arrangement is essentially a recording r-f voltmeter, the voltages in question being those at the input to the receiving set. In particular, the receiver was usually an HRO or an HQ-120-X, the recording mechanism being of the Micromax line-drawing type of recorder or a Brown instrument.

Ionosphere data were collected by means of fixed-frequency pulse transmitters or by means of sweep-frequency pulse transmitters, the returning echoes being displayed on a cathode-ray tube screen, photographs of which were taken at regular intervals. The set up at Stanford University is typical. "Each of the transmitters produced single side-band pulses at the rate of 20 per second. A balanced modulator combined the output of a variable oscillator with that of a 460-kc i-f oscillator and excited a final amplifier which developed a peak power of approximately 1,000 watts. A complete sweep of the range was photographed every half-hour on the screen of a cathode-ray tube which was intensity-modulated by the signal from the receiver. Virtual heights up to about 700 km were recorded."[4,9]

The recorders in the field-intensity work operated from a bridge circuit in the a-v-c circuit of the receiver. Antennas were dipoles properly matched to the transmission lines connecting them to the receiver and oriented to secure maximum signal from the transmitters being recorded. The bridge circuit connecting the HRO receiver to the Micromax recorder (Figure 1) is taken from the final report[7] of Project C-45.

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[b] For a description of the part played by ionosphere prediction and information making possible the choice of best frequencies for radio sky wave propagation over known distances and paths, particularly in setting up the Loran system of long-range navigation aid, see reference 13 of the bibliography.

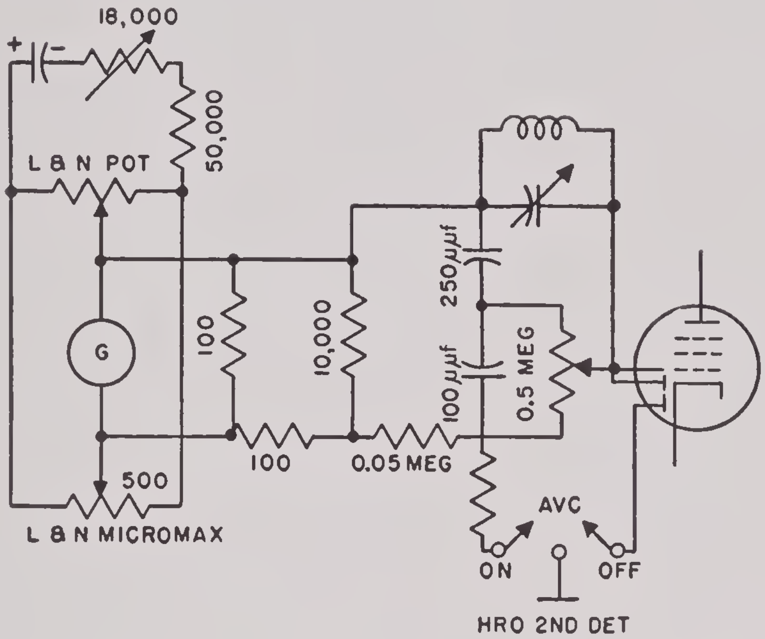

Figure 1. Bridge circuit connecting HRO receiver to Micromax recorder.

More detail on the methods of making echo-height measurements will be found in reference 14 of the bibliography. At the University of Puerto Rico[7] ionosphere recording antennas were vertically directed rhombics, one having 170-ft legs and another having 115-ft legs, the two being connected in series to give reasonably uniform impedance characteristics over the 3- to 12-mc range covered in the measurements. Rhombics were employed in receiving from the stations selected to be monitored for field intensities.

18.3 SOLAR AND GEOMAGNETIC OBSERVATIONS

Under Project C-53[12] a correlation of solar and geomagnetic observations with conditions of the ionosphere was carried out. Weekly reports of magnetic activity were received from the stations of the U. S. Coast and Geodetic Survey at Cheltenham, Maryland; Tucson, Arizona; Sitka, Alaska; Honolulu, Hawaii, and San Juan, Puerto Rico; and from the magnetic observatories of the Department of Terrestrial Magnetism at Watheroo, Western Australia; Huancayo, Peru, and College, Alaska.

Solar observations were conducted for this service at the following places: Climax, Colorado, U. S. Naval, McMath-Hulbert, and Mount Wilson observatories.

A study of the progress of sun-spot cycles and of solar-activity cycles, magnetic disturbances, coronal intensities, sporadic E-region ionization, radio fadeouts, aurora displays, and radio conditions during the partial solar eclipse of February 4, 1943, at College, Alaska, was made under the Projects conducted by the Department of Terrestrial Magnetism.[3,10,12]

18.4 CONCLUSION

The advantages to the Military of having predictions of radio communication effectiveness in planning campaigns are obvious. In 1942 it became important to have certain forecasts some weeks in advance and other types a few days in advance. Radio and ionosphere forecasts of the first kind were prepared and distributed by IRPL. Those of the second kind were prepared for a few months by CIW, and in October 1942, this forecasting function was transferred to IRPL, while the function of centralizing solar and cosmic data for this purpose was retained by the CIW. The final report, Project C2-49,[11] gives an example of the weekly forecast and also of the type of information that was distributed by telephone to proper services. As an example of special warnings, on October 28, 1942, an ionosphere storm was forecast and the information distributed by telephone a few hours in advance, thus warning communications services that trouble might be experienced.

Projects C-9 and C-44, NBS, resulted in radio transmission handbooks for frequencies of 1,000 to 30,000 kc and show conditions under which the different radio frequencies would be useful for the time periods covered, the winter of 1941-42 and the summer of 1942.

PART VII

APPARATUS DESIGN

CONFIDENTIAL

Chapter 19

U-H-F FIELD-INTENSITY MEASURING EQUIPMENT

Development of two field-measuring sets, including receivers, signal generators, and antennas, for the regions 300 to 1,000 and 1,000 to 3,000 mc. The first units of the kind covering these frequencies were developed under this project.

19.1 INTRODUCTION

The object of the research conducted under this project[a] was to develop field-strength measuring equipment for the u-h-f and microwave regions which were coming to be of the utmost importance for radio communication and radar work beginning to be used at the time, early 1941.

At the time there was no commercial field-strength equipment of any kind for use at frequencies above 50 mc. Some experimental instruments had been built for higher frequencies but each design was for a very limited band. The aim of this development was to provide in a single equipment, or in not more than two equipments, the means for making precise field-strength measurements in the range from 300 to 3,000 mc. The equipment had to be as simple as possible for field operation considering the precision with which the measurements were to be made.

The problem was complicated by the fact that at that time there was no receiver or signal generator available for operation across the desired band; in fact, except for some experimental types of radar receivers which are single-frequency devices, there were no receivers at all available for frequencies above a few hundred megacycles. Another serious difficulty was that none of the conventional vacuum tubes would function either as oscillators, amplifiers, or detectors in this range. Tube development, however, was being pushed very hard at the time and some experimental tubes which would perform the required service became available. The development of silicon and iron-pyrite detectors assisted materially by supplementing tubes in some of the circuits.

Under the project, two sets of equipment were designed, built and tested in the field with satisfactory results. A few of the 300- to 1,000-mc instruments were factory-built and proved valuable in testing radar receivers, in addition to their primary job of measuring field strength. Figure 1 shows a complete 300- to 1,000-mc system including the signal generator, receiver, and the associated power supplies. The 1,000- to 3,000-mc units demonstrated in March 1942 were adjudged too complicated for regular production at the time.

19.2 DEVELOPMENT PROBLEMS, 300 TO 1,000 MC

The major problem in the design of the lower-frequency model was the development of an oscillator that would cover continuously the wide frequency range between 300 and 1,000 mc in a single band. There were two aspects of this problem, choice of a suitable tube and design of a suitable continuously variable tuned circuit.

19.2.1 Tuned Circuit

An improved version of a tuned circuit previously described and used in a wavemeter covering the range of 50 to 400 mc was employed in the signal generator. It consisted of an annular stator assembly carrying two sets of 90-degree stator plates mounted diametrically opposite each other within the ring and a rotor carrying fan-shaped plates that mesh with the stator plates. That portion of the ring not carrying stator plates served as a tuned-circuit inductance. Displacement currents passed from one set of stator plates to the other through the rotatable fan-shaped rotor plates and the complete plate assembly served as a continuously adjustable "series-gap" capacitor. When connections were made to two stator-plate assemblies, therefore, the developed impedance was that of an antiresonant circuit whose natural frequency could be varied without the use of sliding contacts by means of an angular shaft rotation.

The mechanical design of a tuned circuit of this type, called a butterfly circuit from the appearance of the rotor (Figure 2), leads to wide tuning range, because the rotor plates act not only as capacitor plates but as short-circuited turns, or eddy-current shields, in the magnetic field. When the rotor plates are turned, they vary both the capacitance and the in-

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[a] Projects C-5 and C-5a, Contracts NDCrc-141 and OEMsr-289. General Radio Co.

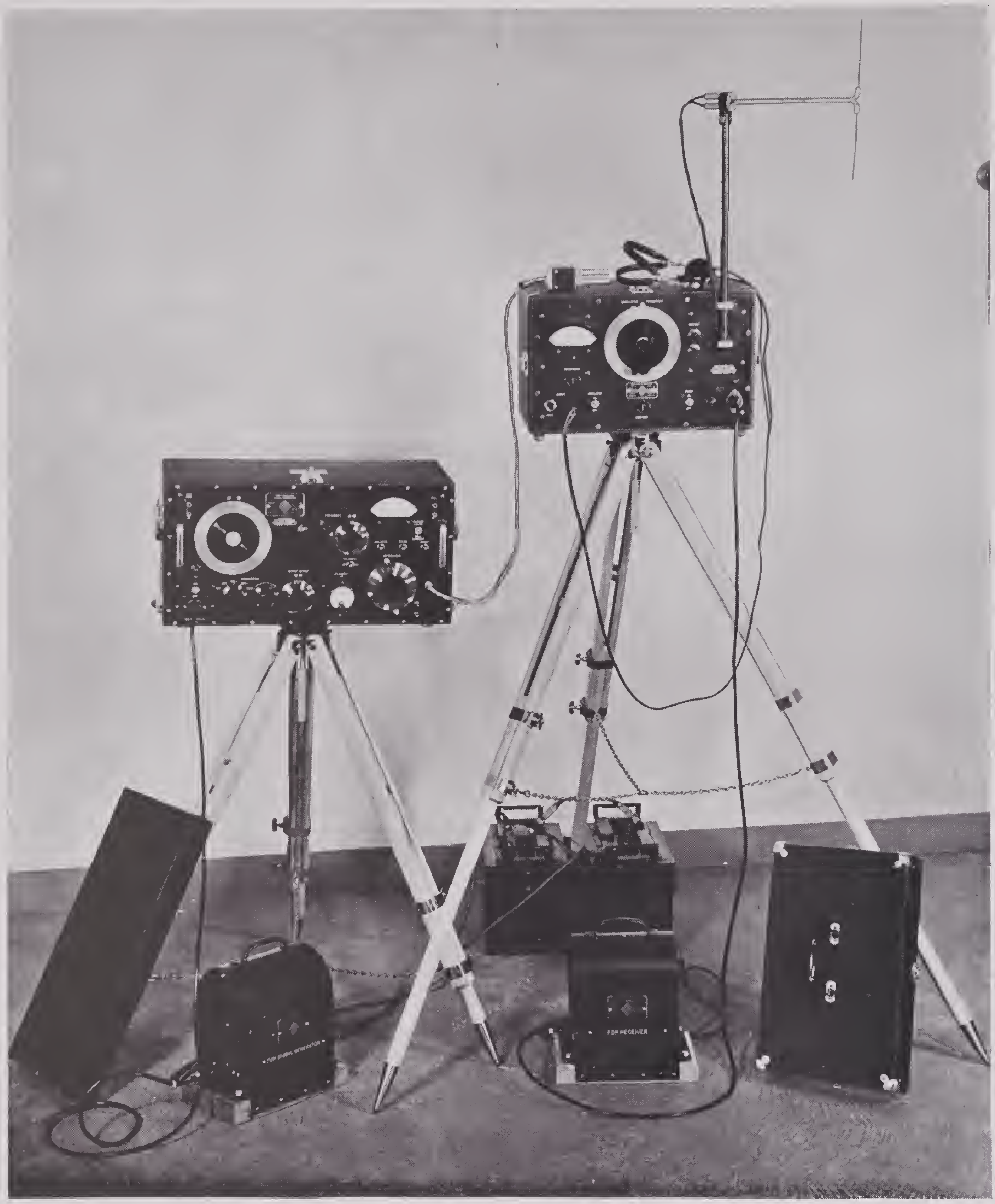

FIGURE 1. Complete setup of the signal generator and receiver with sources of power supply.

ductance. The eddy-current shielding is most effective when the rotor plates are completely out of mesh with the stator and the effective inductance and capacitance are, therefore, both minimum. Conversely they are both maximum when the rotor plates are fully meshed with the stator plates.

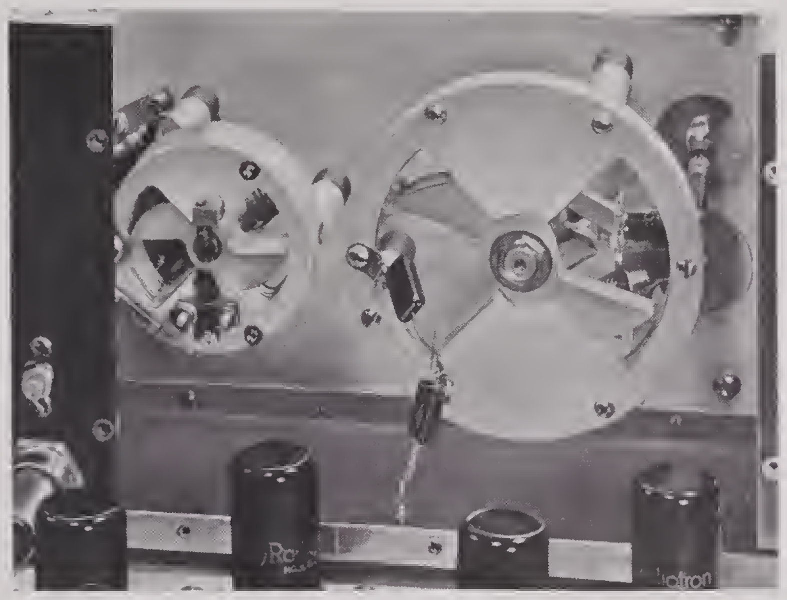

Figure 2. Butterfly circuit used in 300- to 1,000-mc receiver

19.2.2

Tube Choice

At the time, there were few tubes which would work at all much above 500 mc. The particular tuned circuit developed further restricted the tube types that could be employed, the choice rapidly narrowing down to the Western Electric 368-A which was then in commercial production. Western Electric Type D-160127 (1221-Y) coaxial-line tube and General Electric Types ZP-423 and ZP-446 lighthouse tubes were not satisfactory for the butterfly-tuned circuit because of the distribution of interelectrode capacitances, although their ultimate frequency limits with other types of circuits are much higher than the 368-A tube.

19.2.3

Output System

Because a balanced output was desirable and because a very simple design could be evolved, the mutual-inductance type of attenuator was adopted. To give maximum accuracy in standardizing the attenuator output, balanced iron-pyrite crystal rectifiers were mounted immediately at the terminals of a probe in which the output cable terminated. The coupling from the oscillator to the attenuator was obtained by an open-wire pickup lead of variable length that coupled capacitively to the oscillator and that served as the source of magnetic field within the attenuator.

19.2.4

Performance of Butterfly Circuit

The butterfly circuit was found to perform satisfactorily over the range of 300 to 1,000 mc, but at higher frequencies some trouble was encountered in obtaining proper operation.

19.3 1,000- TO 3,000-MC SIGNAL GENERATOR

For this instrument the 368-A tube was used with a single-ended transmission line oscillator circuit in which the fundamental frequency range was from 500 to 1,200 mc, with production of strong second and third harmonics. Means were provided for coupling the transmission line to the attenuator so that the second or third harmonic of the oscillator could be selected at will, the second harmonic being employed to cover the range 1,000 to 2,400 mc and the third harmonic for the range 2,400 to 3,000 mc.

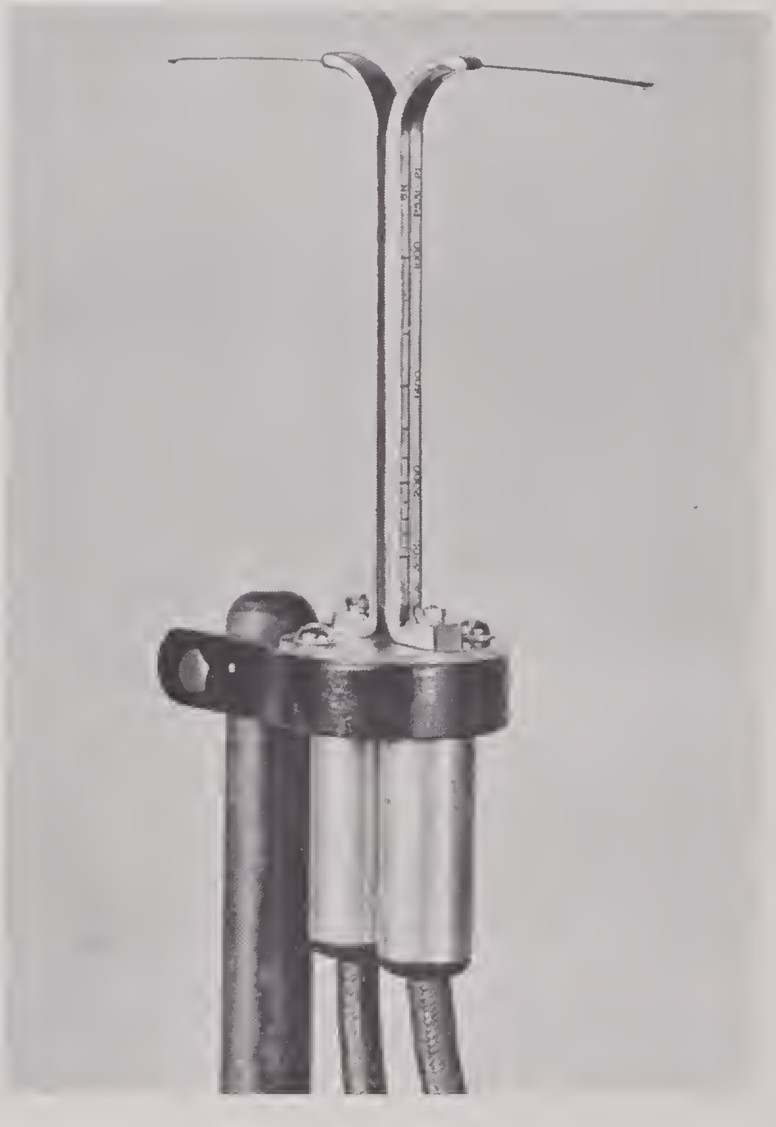

Figure 3. Details of flexible antenna made of steel tape which is extended to correct length.

19.4

RECEIVERS

In the design of the lower-frequency receiver, the oscillator frequency chosen was half the signal-generator frequency. This choice was made to save power and expense, since the Type 955 acorn tube could be used instead of the 368-A. Sufficient power for a heterodyne voltage at the second harmonic could be obtained without difficulty. Antenna tuning was effected by a butterfly circuit. An iron-pyrite crystal acted as a mixer and was connected directly across the antenna-tuning circuit. The five-stage i-f amplifier with cathode-resistor stabilization of input admittance provided a 2-mc band width with center frequency of 30 mc. An input signal of 10 μv could be distinguished through the background noise produced by the crystal mixer and the first i-f stage.

Design of the 1,000- to 3,000-mc receiver (Figure 4) followed the design of the 300- to 1,000-mc instrument. The oscillator covered the range 500 to 1,000 mc, the antenna-tuning butterfly was scaled down three to one, the second harmonic of the oscillator beating with the incoming signal from 1,000 to 2,000 mc and the third harmonic on the range of 1,500 to 3,000 mc.

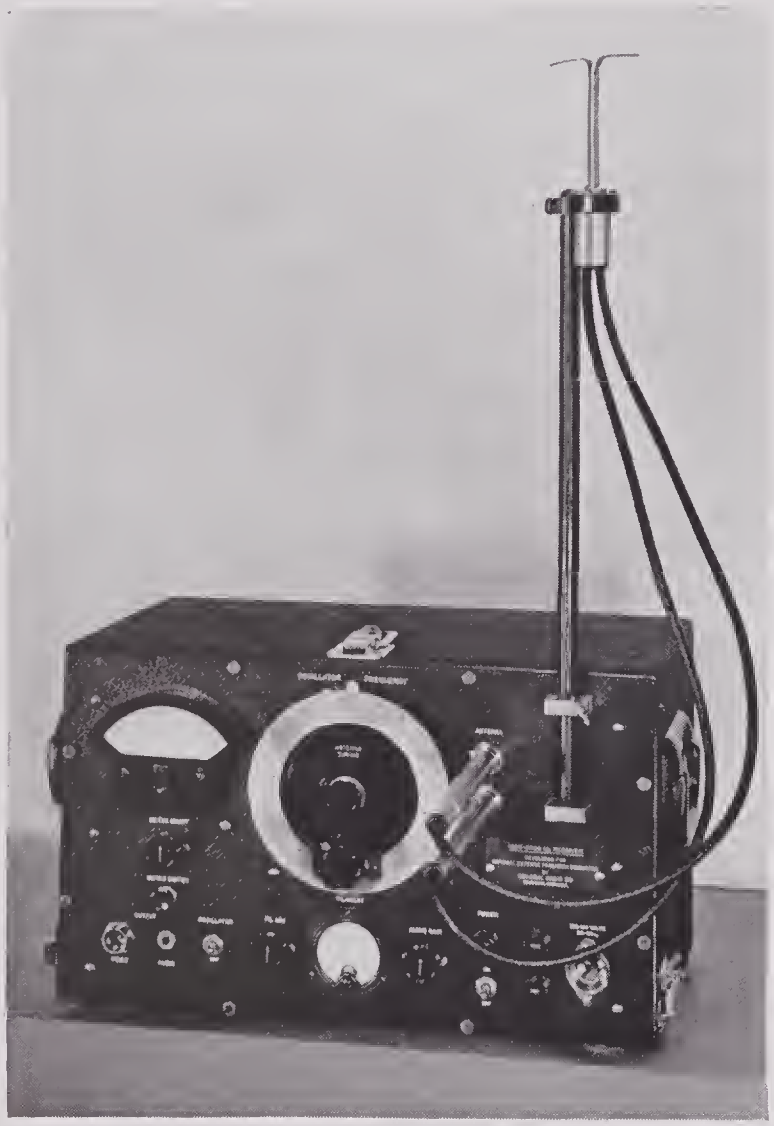

Figure 4. The high-frequency (1,000- to 3,000-mc) receiver.

Chapter 20

AIRCRAFT FACSIMILE SYSTEM

Development of lightweight, compact, high-speed scanner and recorder to be used to test the efficiency of facsimile communication between a military airplane and ground. Synchronization was effected by means of transmitted signals instead of by the customary method of using tuning forks. Transmission rate of 48 sq in. per minute with copy 8 in. wide was effected.

20.1 STATE OF THE ART

At the time this project[a] was started early in 1942, there were no facsimile systems operating faster than about 20 sq in. per minute. All the scanners were of the drum type in which each successive message had to be fastened on the drum and the machine started. Most of the recorders were also of the drum type, using photographic recording or a stylus on paper of the Teledeltos type. Some recorders using continuous carbon paper supplied from a roll were available, but their speed was limited by the mechanical movement of the printers to not more than 12 sq in. per minute. Also available were recorders using a continuous roll feed supplying electrolytic paper, but again the speed was low.

Synchronizing methods in use on low-speed systems depended on fork frequency standards at the two ends of the circuit, the forks driving synchronous motors which in turn rotated the drums. Some work had been done in transmitting a synchronizing signal along with the picture over a radio circuit but circuits using this principle were not in operation. Little or no experience had been had with synchronization of machines running at drum speeds greater than 200 to 300 rpm.

20.2 ACCOMPLISHMENTS OF THE PROJECT

Under Project C-8 much progress was made on mechanical designs of scanner and recorder, in the chemistry of recording solutions for higher speeds, and in synchronizing methods. In the scanner completed under the contract, the optical system rotated and traveled along inside a stationary semicylindrical piece of clear plastic. The message could be easily loaded by simply raising the cover. With scanner amplifier incorporated in the same case the weight of the whole unit was 31 lb.

At the proposed speed of 48 sq in. per minute only a continuous damp-paper electrolytic recorder was feasible. Furthermore the sensitizing solutions which were satisfactory at the lower speeds would not respond at the higher speeds and new solutions were developed. By using a pretreated roll of damp paper, the mechanical construction of the recorder was made very simple and compact, and an arrangement was found which would retain the damp roll in satisfactory condition for a week or more and would still permit almost immediate starting of the machine when desired.

Because the pretreated damp rolls had limited shelf life, a portable paper-treating machine was developed which represented considerable mechanical progress over earlier, bulky laboratory equipment of this nature.

Under the project, much time was spent on solving the synchronizing problem, which was enhanced by the high speed (an increase from 200 to 600 rpm) at which this equipment was to operate. Three distinct synchronizing systems were developed.

The equipment was field-tested by the Signal Corps General Development Laboratory at Fort Monmouth where the scanner was placed in a moving vehicle and messages were sent back to a fixed receiving station. Except for some jitters caused by gearing irregularities the system worked satisfactorily at a speed several times faster than that of previously existing equipment.

About the time the unit was field-tested, the thinking of the military people changed and what was now desired was a combined scanner-recorder with tuning-fork synchronization instead of by transmission of a reference over the circuit. The mechanical design of the compact scanner developed under Project C-8 did not lend itself to direct coupling with a recorder. Field tests, therefore, did not include plane-to-ground transmission.

Thus the original intent of the project, to develop a compact high-speed facsimile system to be tested between airplane and ground as a means of indicating the military value of such communication, was only partially carried out due to change of military plans.

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[a] Project C-8, Contract NDCrc-88, RCA Manufacturing Co.

20.3 SYNCHRONIZING METHODS DEVELOPED

Since tuning-fork synchronizing apparatus developed before World War I was heavy and required considerable power (25 lb, 24 watts) other means of synchronization were sought. Furthermore, if forks were used at scanner and recorder, automatic phasing would be necessary, since hand phasing would be too slow at the operating speed desired and would require constant attention. Thus special phasing signals would be necessary. Synchronizing and phasing become two problems, therefore, and merge into a single problem only in the limiting case where the frame or phase signal also controls recorder speed.

Where synchronism is maintained by a transmitted signal it is still important that the scanner speed be essentially constant. Some slow variation can be allowed, but abrupt change will result in a jog in the copy. To be effective, phasing and "sync" signals must be positive in action; that is, no change in synchronization or phasing should occur except as such change is initiated by the transmitted signals. In this manner, fading or drop-out of signal will not cause lack of synchronization or phase. This requires that the two signals be separated far enough from the picture signals so they can be separated by filters.

Assuming that the sync tone is received without distortion, the problem remains of controlling the recorder motor speed from this tone. At the speeds of transmission required, the sync tone should be 300 cycles or higher and this would require a 300-cycle synchronous motor, a product not available at the time of the project. It would be more desirable to use a standard 60-cycle motor, provided means could be found to limit the phase shift with changes in voltage and load.

The research on synchronizing methods, therefore, was devoted to means for controlling a 60-cycle motor from various types of transmitted signals.

20.3.1 Synchronizing Method No. 1

In the first method investigated, sync and phasing were transmitted as separate signals, the former being a steady tone of 960 cycles generated by a tone wheel on the scanner motor and the second a 4-kc signal at the start of each scanning line. At the recorder, the two signals were separated by filters. The incoming 960-cycle signal was used to lock in a 960-cycle phase-shift oscillator, which then provided a reference tone representing the scanner speed. It was stable enough to hold approximate control even when no 960-cycle tone was being received from the scanner. A second 960-cycle oscillator was locked in on a tone wheel on the recorder shaft and provided the reference tone representing the recorder speed. These two tones were combined and rectified, the resulting current being zero when the phase difference was 180 degrees. Maximum current occurred when the two systems were in phase. This direct current could be employed to control the recorder motor speed in the following manner.

The 60-cycle vibrator driving the recorder motor was self-excited at approximately the correct frequency, but by varying the coil excitation voltage the frequency could be changed over a range of ±2 cycles. The rectified beat between the reference tones controlled a load tube across the vibrator excitation and therefore controlled the vibrator frequency. Normal setting was made with the two tones exactly 90 degrees apart (center of the range) and the vibrator at exactly 60 cycles. If the recorder speeded up slightly, the phase shift increased and excitation of the vibrator was reduced by the increased loading so that the vibrator frequency was reduced.

This method was capable of holding the phase displacement at ±6.53 degrees of the 60-cycle supply.

Line phasing was accomplished by pulsing the load tube to the full slow position and beyond control of the phase corrector each time a 4-kc signal was received. When the recorder was jogged out of sync a sufficient number of times to come into correct phase position, a commutator on the recorder shaft shorted out any further action of the phasing pulses.

This system had an important ability to work with poor received signals. The filtering action of the 960-cycle RC oscillators was so good that no other filter was required to separate the 960-cycle tone from the other signals. Phasing was fast and accurate, only a few scanning lines being lost at the start of a message.

The method, however, proved to be too precise; hunting easily developed. To remedy this trouble, tightness of control was halved by dropping the sync tone to 480 cycles. This improved the action but the system was still too sensitive.

20.3.2 Synchronizing Method No. 2

In this method the sync reference was taken from the phasing signal once per scanned line. This method would require no special sync frequency in the transmitted signal, filtering in the recorder would be simplified, and the full modulation capacity of the radio transmitter could be used for the picture signal.

In this method the vibrator was tube-driven instead of being self-excited, the driver tube receiving its excitation from a 60-cycle oscillator of the RC phase-shift type and adjustable over a small frequency range. The phase signal of one pulse per scanning line was separated from the picture signals by a limiter-filter arrangement, rectified and shaped to approximate a half cycle of a 60-cycle wave. This shaped pulse was fed to the 60-cycle oscillator and locked it into synchronism on the nearest half cycle. Accurate synchronism was held by this single locking pulse once per scanning line (every 6 cycles).

Phasing was accomplished by having the phase pulse change the oscillator frequency to 58 cycles instead of 60 whenever the recorder was out of phase. This was done by changing the network timing by commutator, so that the frequency change was effected instead of pulsing the RC oscillator. This caused a rapid drift of the recorder to a new phase position at which point the oscillator was immediately restored to 60 cycles and pulsed into synchronism.

In normal operation this system was very accurate and phasing was even more rapid than in the first system. However, two commutators were required on the recorder and because of gradual changes due to brush wear, etc., they did not maintain their proper relation to each other over a long period. This caused jitter. Furthermore, the 60-cycle RC oscillator could shift in frequency more rapidly than the vibrator and motor could follow. The result was that heavy surge currents would pit and burn the contacts so that the vibrator was practically worn out after a few hundred hours of operation.

20.3.3 Synchronizing Method No. 3

A modification of the second system was finally adopted. In this method, the RC oscillator driving the vibrator was replaced with an LC oscillator which could not change its frequency so rapidly as the vibrator and much better control was experienced. The vibrator had a normal life in this system. The control system for correcting the oscillator frequency was changed to a self-balancing bridge instead of depending on pulsing to a new frequency with each successive scanning line. With this bridge no action took place if no phasing signal was received, the oscillator thereby holding the frequency to the value to which it was last set.

Rapid drift on fade-out of the signal was thereby prevented. Only one commutator was required and this did not have to be phased in with the 60-cycle motor. Brush wear, therefore, did not disturb synchronism. A complete description of this system and its wiring diagram are to be found in the final report[1] of the project.

Chapter 21

ULTRA-HIGH-SPEED FLASH TELEGRAPHY

Magnetic tape driven at slow speed has impressed upon it a 20-word code message recorded at a rate of 30 words per minute; this 40-second message is then transmitted by running the tape at 100 times the recording rate. At the receiving end of the circuit the message is recorded at high speed and transcribed at low speed. Transmission time is 0.4 second or less. Transmission speeds as high as 3,000 to 9,000 words per minute are possible. The essentials of the completed units and the requirements for the radio circuits are included in this summary. The final report,[1] from which this summary is condensed, gives more details of solutions to difficulties that occurred in development, also numerous oscillographs of test messages sent over the system.

21.1 INTRODUCTION

The objectives of this project[a] were to simplify code sending and receiving equipment, which usually consists of perforated tape transmitters and oscillographic or other recorders, and to achieve telegraphic speeds of the order of 3,000 words per minute. Flash telegraphy as a means of radio communication appears to have very definite advantages when security from direction-finding determination and interception is desirable. By use of special limiting-type amplifiers and modulators, discrimination against static and multiple-path transmission was obtained.

Any method of flash telegraphy requires wider r-f bands and imposes stricter requirements on signal-to-noise ratios and selective fading than does transmission and reception at manual speeds. Although the Services did not adopt the flash system, largely because of limitations in the then existing radio equipment, the limitations could have been reduced substantially by engineering a complete flasher system, including not only the terminal units but radio transmitter and receiver equipment designed to meet the special requirements of flash telegraphy.

The achievement of a speed-up factor of some 100 to 1 and telegraphic speeds of 3,000 words per minute seems to be unique. The need for maintaining radio silence during naval operations and the great desirability of security in the field of combat indicate that the terminal equipment and methods described here offer distinct answers to these security problems.

21.2 FUNCTIONAL OPERATIONS OF THE FLASHER

The method of magnetic-tape recording employed involves essentially the transverse magnetization of a thin ribbon of magnetic tape. Ordinarily, one of two recording procedures may be used.

D-C Erase. In this method the tape is erased by a d-c flux sufficient to saturate the tape. Hence, as the tape leaves the erasing pole piece it is magnetized in one direction. In recording, the signal or a-c flux is superimposed on a d-c bias flux having a sign opposite that of the d-c erasing flux. This is done in such a way that the signal flux increases or decreases the bias flux over a linear range of magnetization. As the tape leaves the recording pole piece, it is magnetized in one direction in varying amounts, in accordance with the signal flux variations.

A-C Erase. In this method, the tape is erased by an alternating flux having a frequency that is high compared with any signal frequency and a wavelength on the tape that is small compared with the pole piece dimensions. As the tape leaves the pole piece, it is completely demagnetized. In recording, the signal flux is superimposed on a similar high-frequency bias flux. This is done in such a way that the magnetization passes linearly from positive to negative values in accordance with the signal flux. This second method gives a signal-to-noise ratio some 10 db better than that obtained with d-c erase and bias, but requires the use of a high-frequency oscillator. Direct-current bias may also be employed in the case where the erasing is done with high frequency. This gives essentially the same magnetization pattern for a given signal, as is obtained with d-c erase and d-c bias.

21.3 RECORDING CONSIDERATIONS

In the preliminary studies, two methods of recording the telegraph signals were considered. One method involved the recording of the d-c telegraph signals directly, which consist of current intervals corresponding to the dot and dash marks and no-current intervals corresponding to the spaces. The other involved recording an a-c carrier wave as modulated by the d-c telegraph wave, the modulator being of the balanced type so that the modulating or d-c telegraph wave is balanced out. In this case, the signals to be recorded consist of intervals of alternating current corresponding to dot and dash marks and intervals of no current corresponding to the spaces. For low-speed recording, such a signal is obtained by simply keying an alternating rather than a direct current.

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[a] Project C-28, Contract OEMsr-50, Bell Telephone Laboratories, Inc., Western Electric Co., Inc.

D-C Recording

At a telegraph word rate of 30 words per minute, the dot space interval corresponds to the period of a 12.5-cycle wave. A sequence of d-c dots and spaces may be represented as a wave composed of a d-c component, a component of 12.5 cycles and odd harmonic components of this frequency having amplitudes diminishing inversely as the number of the component. Ordinarily, three of the components are sufficient to define the waveform with enough accuracy for telegraph purposes. Hence a frequency range from 0 to some 37.5 cycles is needed for low-speed recording and reproducing.

Recording the d-c telegraph pulses presents no problem and may be done by employing d-c erase and d-c bias so arranged that the d-c telegraph pulses oppose the bias. Thus, the magnetized elements of the tape will correspond to current intervals of the signal. The voltage developed by the reproducer, however, depends on the *derivative* of the tape magnetization, so that significant voltage is obtained only at the beginning and ending of the original current intervals. Reproduced d-c intervals may be obtained by employing a pair of thyratrons or trigger tubes so arranged that voltage of one sign strikes an arc in one tube and extinguishes an arc in the second and voltage of opposite sign strikes the second tube and extinguishes the first.

This method of recording and reproducing was tried in the preliminary studies and found to be feasible. It was felt, however, that the method would be particularly vulnerable to extraneous transient voltages striking arcs in the tubes in random fashion, and the method was abandoned in favor of the one involving the recording of a-c pulses. As the work has progressed, this decision has been questioned, and it is believed now that the first method might well be investigated further, particularly if higher speeds are of interest. The d-c method has the advantage of requiring about half the frequency range required for the a-c method, as will be seen in the following paragraphs.

21.3.1

A-C Recording

When a-c telegraph pulses are to be recorded, the low tape speed is set by the frequency of the a-c wave needed to define a dot interval. At a 30-words-per-minute rate, a frequency of 60 cycles was considered the lowest that could be used. This corresponds to about half of the lowest frequency used in commercial carrier telegraphy. The dot interval of 0.04 second is represented by about 2½ cycles of the carrier wave.

A sequence of 60-cycle dots and spaces may be represented as a wave composed of a 60-cycle component and odd order summation and difference components having frequencies of 72.5, 47.5, 97.5, 22.5, etc., cycles, of amplitudes diminishing inversely as the orders of the components. Ordinarily, three components are sufficient to define the envelope, so that a frequency range from 22.5 to 97.5 cycles or a band width of some 75 cycles is needed. It was found that a tape speed of ½ in. per second afforded, with a little equalization, a good response over the frequency range from 25 to some 200 cycles, and this speed was adopted for low-speed recording.

As the tape speed is increased, the frequency range increases. A high tape speed of 50 in. per second was chosen, which affords a good response up to frequencies of about 10,000 cycles and results in a speed-up factor of 100 to 1. At this speed, the signal frequencies cover the range from 2,250 to 9,750 cycles, or a band width of 7,500 cycles. The 60-cycle carrier frequency appears as 6,000 cycles.

21.3.2

Teletypewriter and Perforated Tape Methods

During the course of the preliminary studies, consideration was given to several operational procedures for the flasher equipment. One of the early procedures considered was that of using teletypewriters for the low-speed keying and transcribing operations, and experimental work was carried out with such equipment. This procedure was abandoned because its use appeared to require a higher degree of synchronization between sending and receiving flasher units than was desirable from a practical standpoint. The use of perforated paper tape for keying the sender and a Boehme ink recorder for recording the signals reproduced by the receiver was considered also. Finally, the low-speed word rate of 30 words per minute, making manual keying and aural reception possible, was adopted. This procedure obviated the need of any close synchronization between sender and receiver, although at that time a starting pulse was considered necessary to receive high-speed flash messages in a fraction of a second.

21.3.3 Starting-Pulse Problems

Several methods employing a starting pulse for receiving high-speed messages were considered. In principle, they involved a connection to the recording coil through a pair of commutator rings. A brush connecting the rings would be released by the starting pulse and would make one revolution only over the commutator, thus permitting a recording to be made during one revolution. The time of one revolution would correspond to the high-speed transmission time. Consideration was given also to the use of a coded type of starting pulse so that the receiver might not be triggered off by a static crash or other random disturbance.

The reason for proposing the use of a starting pulse was to avoid the erasing of the recorded message by the recording bias current. Initially, bias current was considered to be essential, and to avoid erasing it was necessary to disconnect the bias at the end of the high-speed message. The purpose of a bias current is to confine the magnetization to a linear range and thus avoid nonlinear distortion of the signal. Subsequently, it was realized that many of the distortion products of the high-speed signals would fall outside

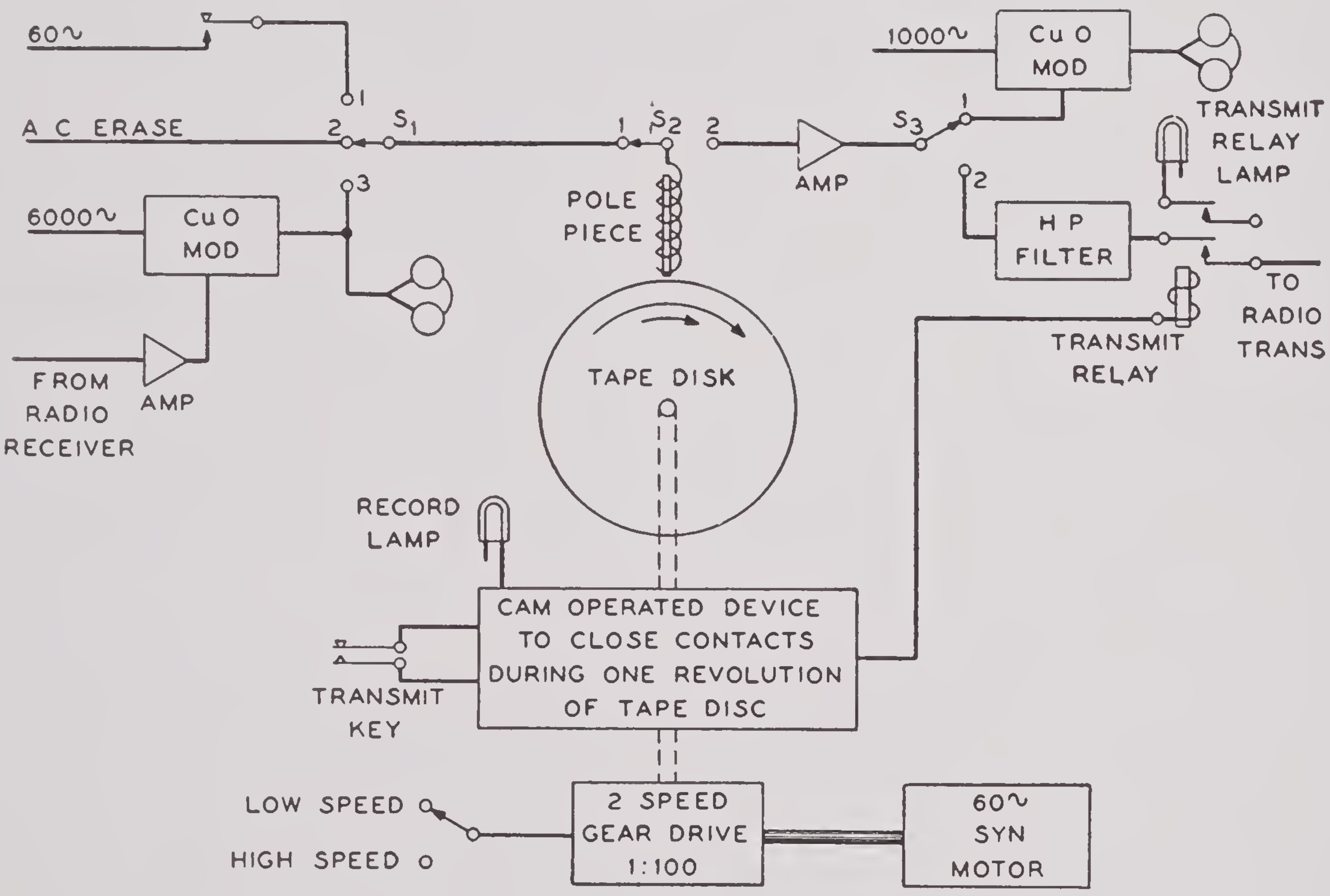


Figure 1. Functional operations of flasher system of high-speed telegraphy.

1. Recording message for transmitting (low-speed operation). Switch S_1 is thrown to position 2 and tape erased for at least one revolution. Switches S_1 and S_2 are then thrown to position 1. Cam-operated device dims record lamp momentarily, thus indicating beginning of time interval of 46 seconds during which messages may be recorded.
2. Confirming recorded message (low-speed operation). Switches S_2 and S_3 are thrown to positions 2 and 1, respectively. Recorded message is picked up, amplified, and used to modulate or key a 1,000-cycle tone to produce audible signals in telephone receivers.
3. Transmitting message (high-speed operation). Switches S_2 and S_3 are thrown to position 2. Recorded message is picked up, amplified, and sent on line to radio transmitter. Line should be capable of transmitting a band of from 2,250 to 9,750 cycles per second. When transmit key is depressed, cam-operated device is energized and acts to close transmit relay for one revolution only of tape disk. Closing of transmit relay is indicated by transmit relay lamp. After lamp flashes, operator may release transmit key.
4. Receiving message (high-speed operation). With tape erased, switches S_1 and S_2 are thrown to positions 3 and 1, respectively. Operator monitors incoming signals and throws switch S_2 to position 2 after message is received. The 6,000 cycle tone is modulated or keyed by rectified output of radio receiver detector circuit. Amplifier connecting receiver and modulator is provided as separate unit, so that it may be located near radio receiver. Line from amplifier to modulator should pass band from 0 to 3,750 cycles.
5. Transcribing received message. Operation same as 2 above.

the frequency range of the tape, and that a bias current would probably not be necessary. Recording without bias was tried and appeared to give satisfactory reproduction, so the procedure of employing a starting pulse was abandoned.

Recording without bias simplified the high-speed receiving operation considerably. It is only necessary to erase the tape clean and wait for the flash message. When it arrives, it is recorded and remains on the tape. A manual operation is required to erase it. The recording process is made highly discriminatory against static, multiple-path transmission, and attenuation or fading by employing a sharply limiting amplifier. It is only necessary to set the gain ahead of the recorder so that static and signals from unwanted paths are just below the recording level. The wanted signal, if it is only a few db or 40 db above this point, will be recorded successfully. On this basis the functional operations indicated in Figure 1 were adopted. The tape disk consists of a loop of magnetic tape mounted on the periphery of a light wheel, and having a length sufficient for recording 22 words at the rate of 30 words per minute.

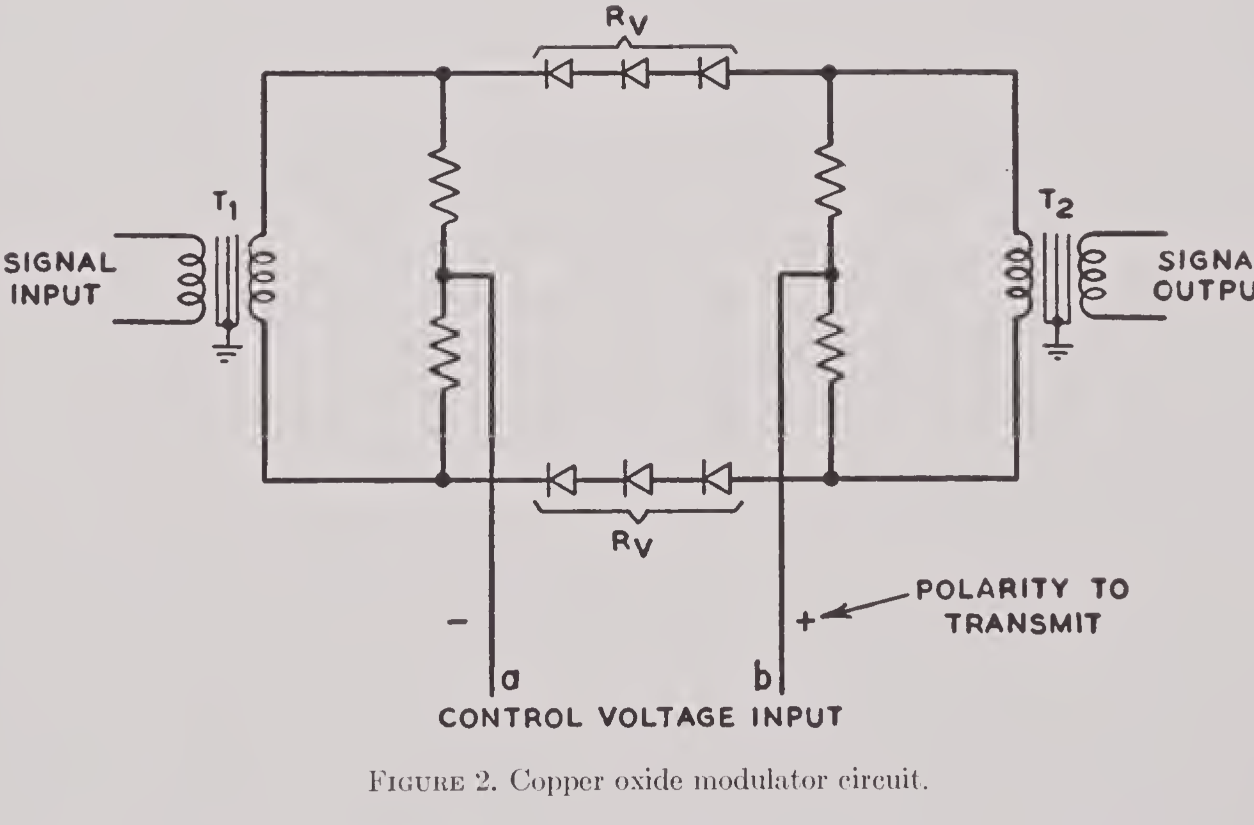


Figure 2. Copper oxide modulator circuit.

21.4 DESIGN OF FLASHER

Proposed design elements having been decided on, the physical design of the equipment as a whole was started. At the same time, detailed studies of the performance of certain of the elements were begun. One such element given detailed study was the copper oxide modulator circuits, the schematic of which is given in Figure 2. The function of these circuits as a variable attentuator is as follows: Each series arm (R_v) is a stack-up of copper oxide disks having unilateral conductivity. A bias voltage is applied at points *a* and *b* of such polarity as to make the transmission path between T_1 and T_2 a high resistance. The d-c signal voltages are applied at *a* and *b* with the opposite polarity, making the resistance low. Thus the device acts as a variable attenuator whose loss is governed by the signal voltage opposing the fixed bias. The change of level obtained is in excess of 40 db. Performance curves are shown in Figures 3 and 4. Further tests were also undertaken with regard to the response characteristics of the tape at the low and high speeds, as existing quantitative data were meager.

Several features of the circuit design were to follow rather conventional lines, but some presented novel problems. Some of these novel features were recording on the magnetic tape without bias; a circuit for transmitting the message once only (or during one revolution of the tape disk; an amplifier to take the detector output of the radio receiver and operate a copper oxide modulator at its output; means for keying a 1,000-cycle tone to the headphones, corresponding to the message recorded on the tape at low speed; means for driving the tape wheel at two uniform rates of speed in a ratio of about 1/100 and for shifting rapidly between these two speeds. One additional feature provided was the incorporation of circuit switching and gear shifting from high to low speed in one control. This control has five positions providing two

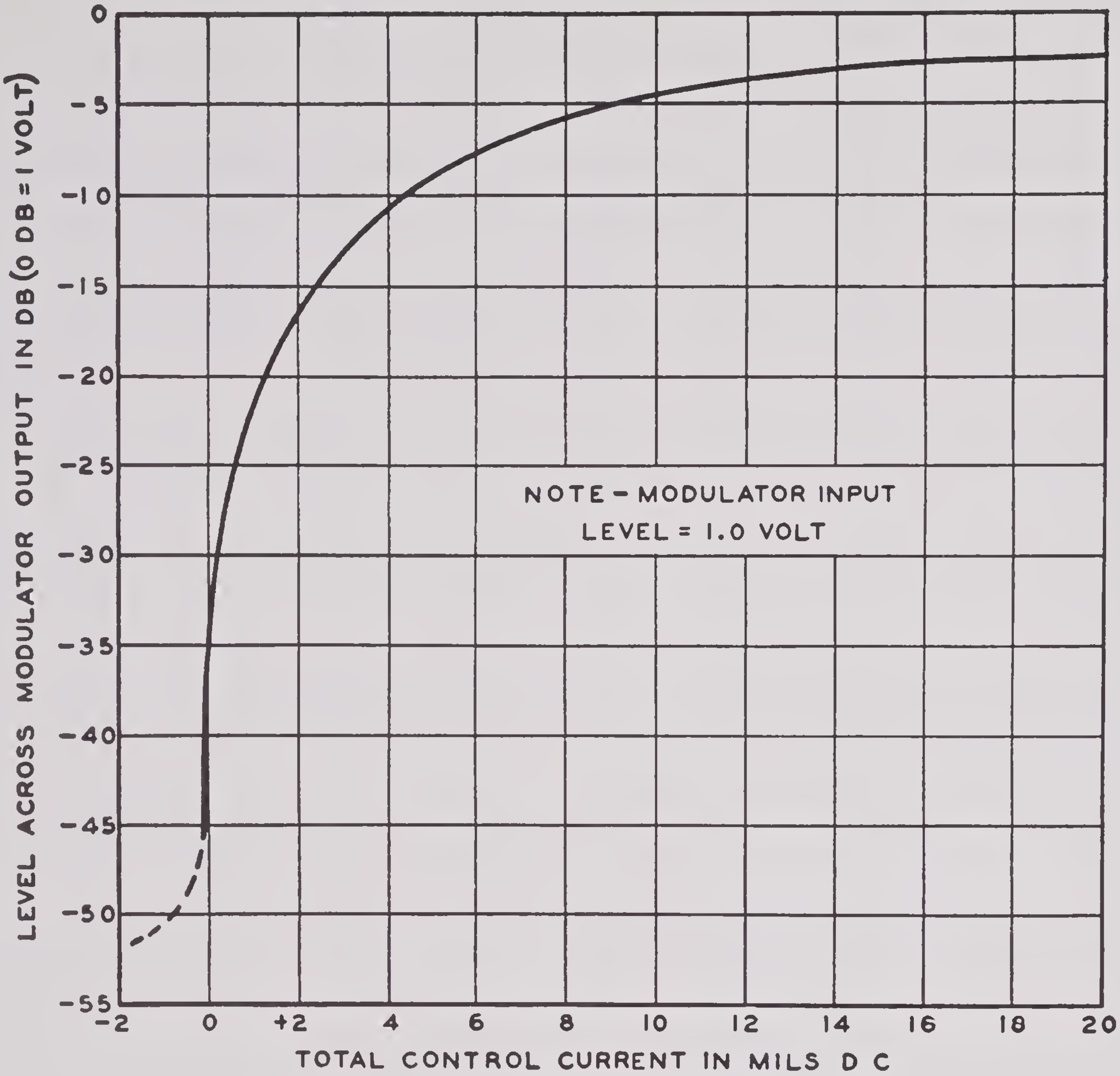


Figure 3. Suppression characteristics of flasher modulator.

circuit conditions at high speed, neutral, and two circuit conditions at low speed.

21.4.1 Tape Drive Mechanism

The high-speed drive consists of a worm on the motor shaft and a worm wheel concentric with tape-wheel shaft. The low-speed drive consists of a train of gears comprising three steps, steps (1) and (2) being spiral gears, and step (3) a worm and worm wheel concentric with the tape-wheel shaft. All gearing is in constant mesh, and the tape-wheel speed change is accomplished by means of a sawtooth clutch which couples the tape wheel to either the high- or low-speed worm wheels. Some care had to be taken to reduce flutter from the gear drive and also, because of the low voltage developed by the reproducing pole piece at low tape speeds, attention had to be given to reducing vibration.

21.4.2 Amplifier for Use with Radio Receiver

This booster amplifier, Figure 5, is a two-stage, high-gain unit of the limiting type. A minimum input voltage of about 0.3 volt (or —10 db per volt) is required for full output, but as the input voltage is increased beyond this value, little change in the output results. Thus a signal of substantially constant amplitude will be furnished to the input of the flasher regardless of variation of the level at the radio receiver output, provided it always exceeds 0.3 volt. In this way, the effects of fading in the radio transmission circuit are minimized.

A gain control is provided at the input of the booster amplifier to reduce its sensiitvity in order to discriminate against interference. The output tube is biased to plate-current cutoff. Under these conditions the output current versus input voltage is nonlinear.

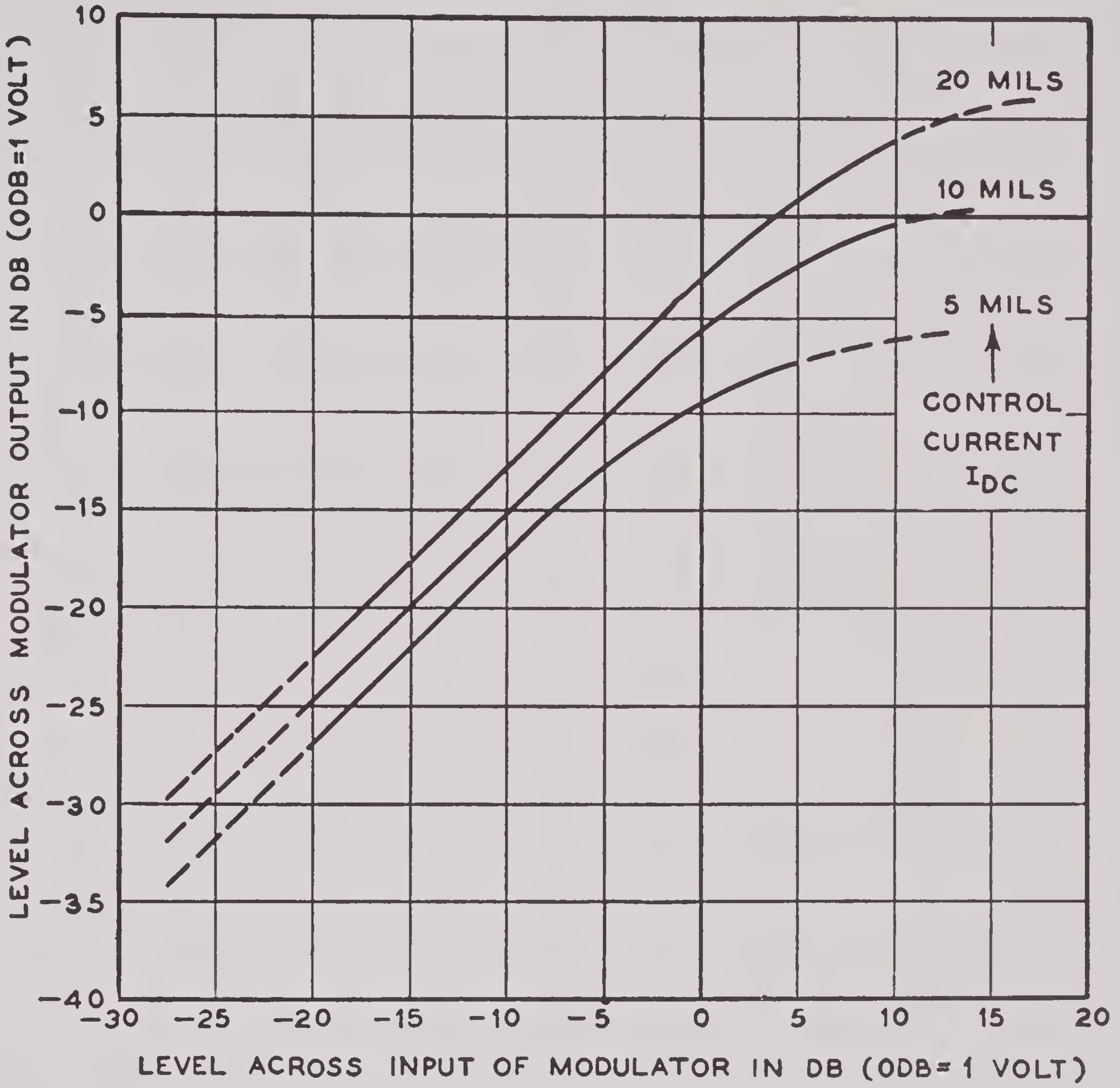


Figure 4. Load characteristics of modulator.

Small input voltages produce only very small plate currents, but input voltages of the order of the bias voltage produce plate currents which are proportionally much greater. In other words, a 6-db increase in input voltage, above some small value, will increase the plate current by 15 db or better. Thus, by properly adjusting the gain, a signal 6 db above the noise level prevailing over the recording period will, by virtue of this expansion, give satisfactory reception.

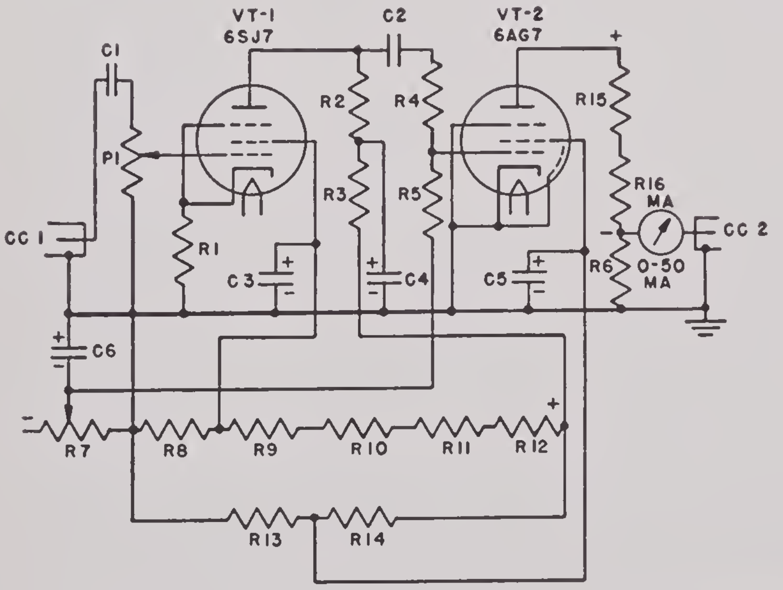


Figure 5. Booster-amplifier limiter circuit.

21.4.3 First Trials as a Complete System

After making volume runs and other tests on the completed machines, they were interconnected by a wire line through a vacuum-tube rectifier of the biased type. This was used to simulate the entire radio link, including keyer, transmitter, and receiver. The output of this rectifier was connected to the input of the booster amplifier. A series of dots was recorded on the tape of the sending flasher by means of a commutator and test transmissions were made. The signal was accurately received, however, over only a very small volume range, about 8 db, rather than over a range of 30 to 40 db as had been anticipated. Particularly unaccountable was a filling-in of the received signal at high input levels.

21.4.4 Distortion from Tape Recording

Oscillograms of the reproduction from the tape at low speed revealed that if pulses of uniform length of time were recorded, nonuniform pulses were reproduced. Increase of the pulse length occurred due to transients of considerable amplitude, while shortening was also observed. This was due to recording with d-c erase and no bias when receiving signals at high speed.

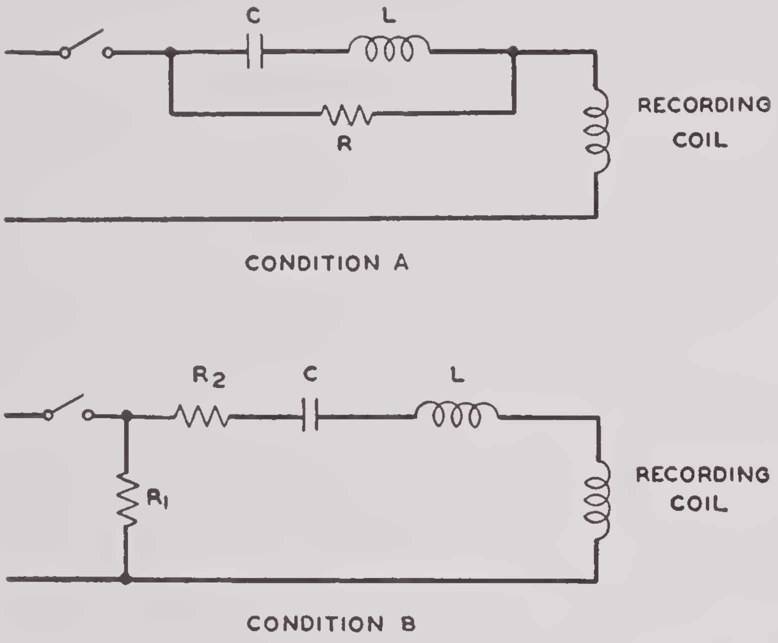


Figure 6. Pre-equalizers for magnetic tape recording.

Under these conditions, the tape is magnetically saturated by the erase current. The received signal then can change the magnetization in one direction only. In other words, the tape acts as a rectifier in the same manner as a vacuum tube biased to cutoff. Under these conditions it is possible to incur a considerable shortening of the pulse length, depending on the phase of the received signal. For example, if the received signal corresponding to a dot happened to consist of 2½ cycles of the 6,000-cycle carrier, the wave will have 2 half-cycles in one direction and 3 half-cycles in the other. Thus the recorded wave might be shortened 40 per cent if the initial and final half-cycles had the same polarity as the erase current. The remedy for this distortion was to employ a-c erase so as to leave the tape nonmagnetized. When this was done using a 30-kc erase frequency, pulse shortening was eliminated.

Pulse lengthening due to transients was reduced by the following prodecures. Care was taken to insure that the locally generated 60- and 6,000-cycle carrier waves and the 30-kc erase wave were free from distortion. Suitable frequency equalization and control of the input levels to the recorder effected further improvements.

In particular one change helped reduce the transients materially. Low-speed keying had been done simply by opening and closing the series circuit at the input to the recording pre-equalizer. It was found that shunting the equalizer input with a resistance and breaking ahead of this point made considerable improvement (see Figure 6). Condition *A* shows the original circuit and condition *B* the improved one. The frequency response obtained off the tape using

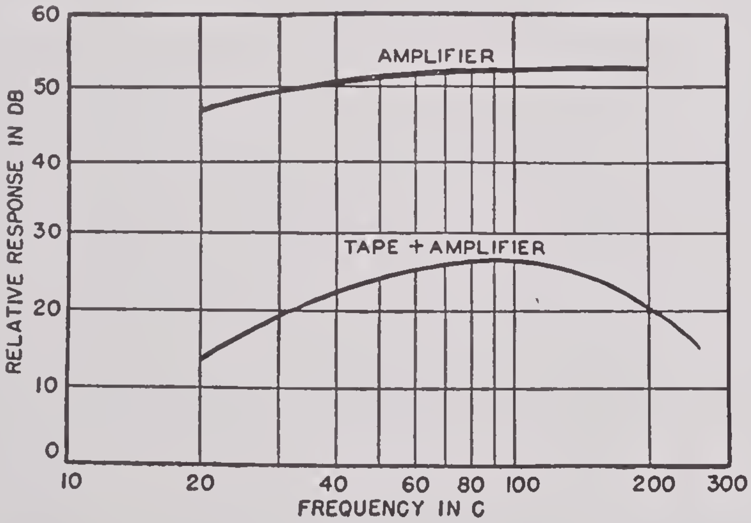


Figure 7. Relative response of magnetic tape at speed of one-half in. per second.

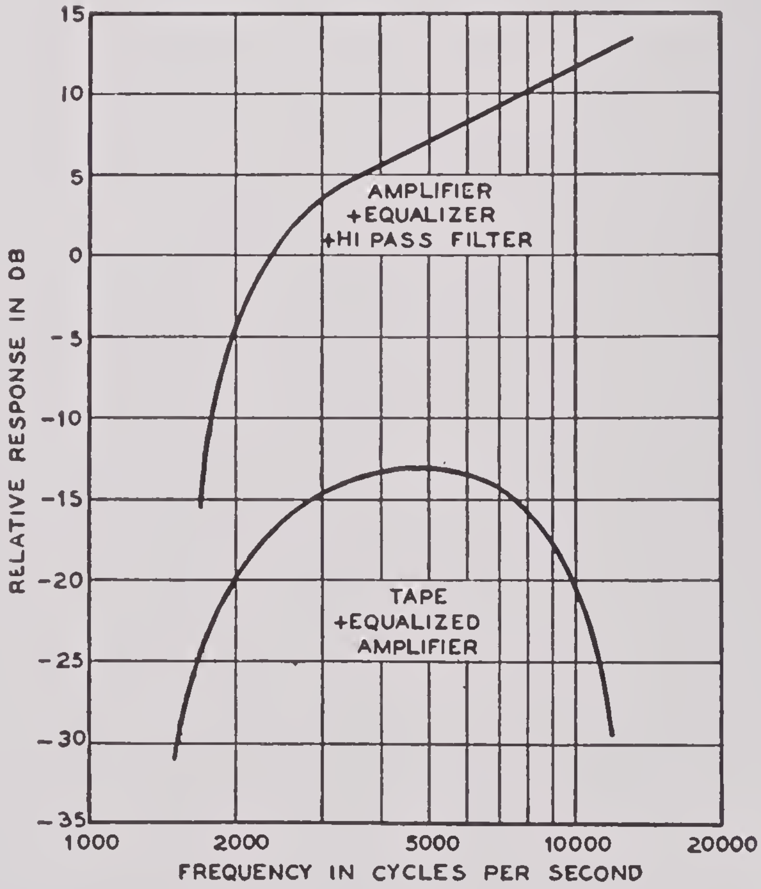


Figure 8. Relative response of tape at speed of 50 in. per second.

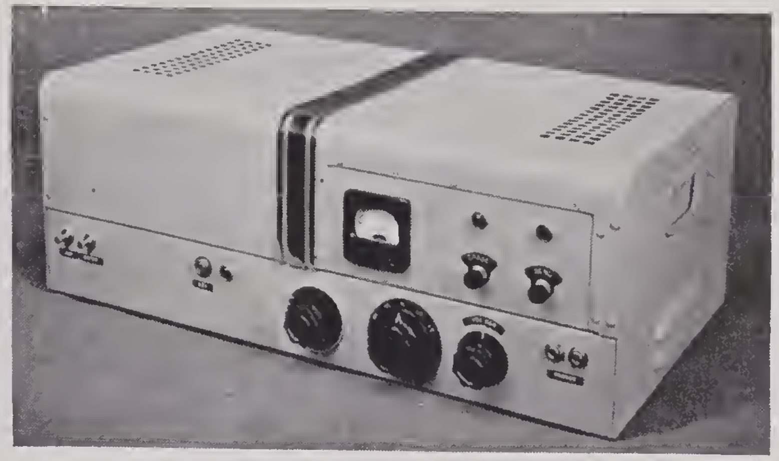


Figure 9. Flasher unit complete.

this pre-equalizer is shown by the lower curve of Figure 7. The upper curve gives the response of the reproducing amplifier, which was unequalized.

The frequency response at high speed was not so good, relatively, as at the low speed. Post-equalization, or, in other words, equalization of the signal reproduced from the tape was necessary in addition to the pre-equalization used. The response obtained under these conditions is shown by the lower curve of Figure 8, while the upper curve shows the characteristics of the equilized amplifier plus a high-pass filter which was inserted in the output. This filter was found helpful in discriminating against low-frequency noise of a microphonic nature originating in the amplifier and caused primarily by the impact of the timing cams on the timing cam contact springs, which occurs once per revolution of the tape disk.

With these improvements and modifications, the total maximum distortion in the complete process of transmission from one terminal to the other, including final low-speed reproduction, was reduced from a possible 80 per cent to a possible 25 per cent. Satisfactory transmission was now obtained over a range of about 35 db at the input to the booster amplifier. It was found that the output-level adjustment of the sending machine was somewhat critical, as had been

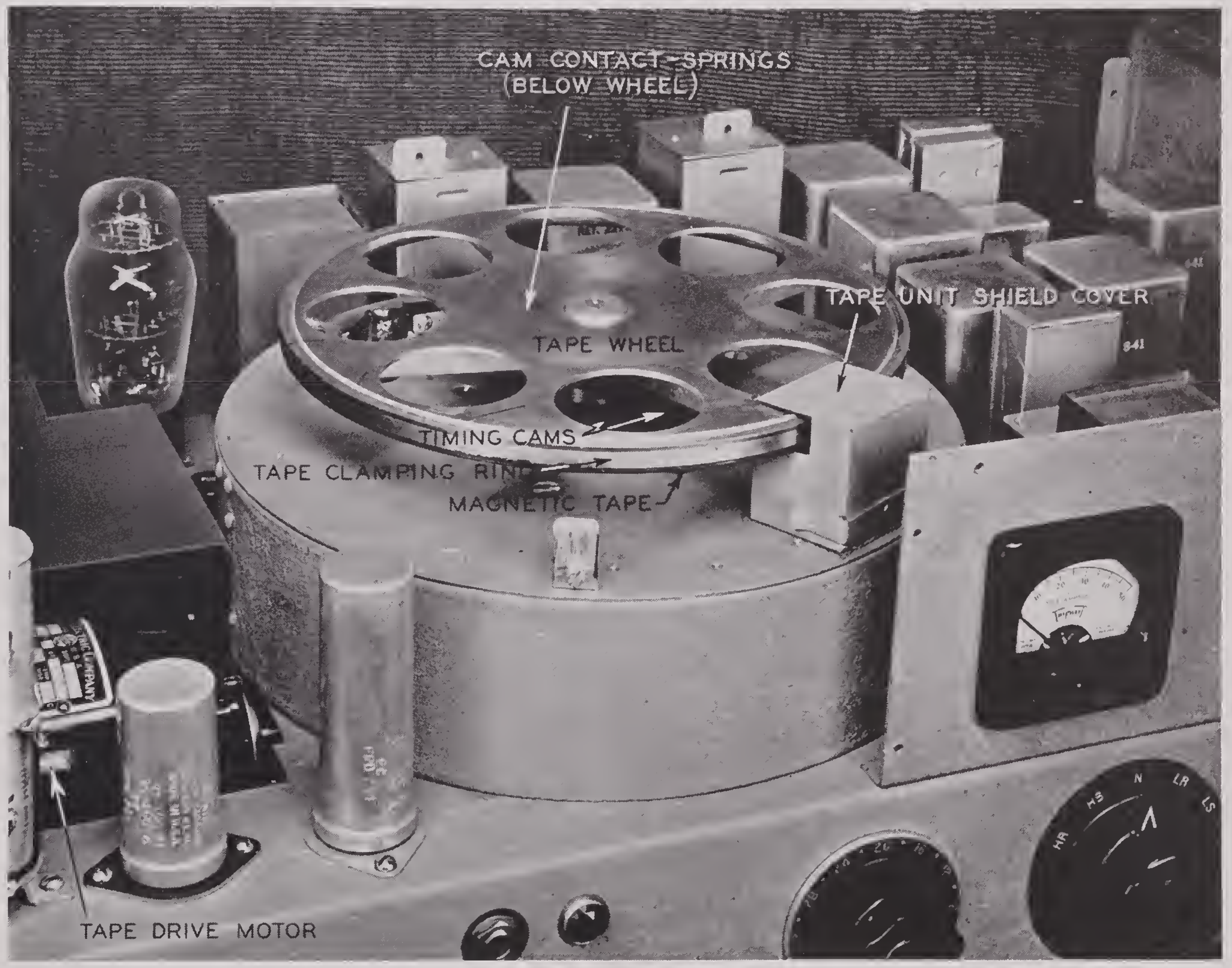


Figure 10. Flasher with cover removed showing position of tape wheel, etc.

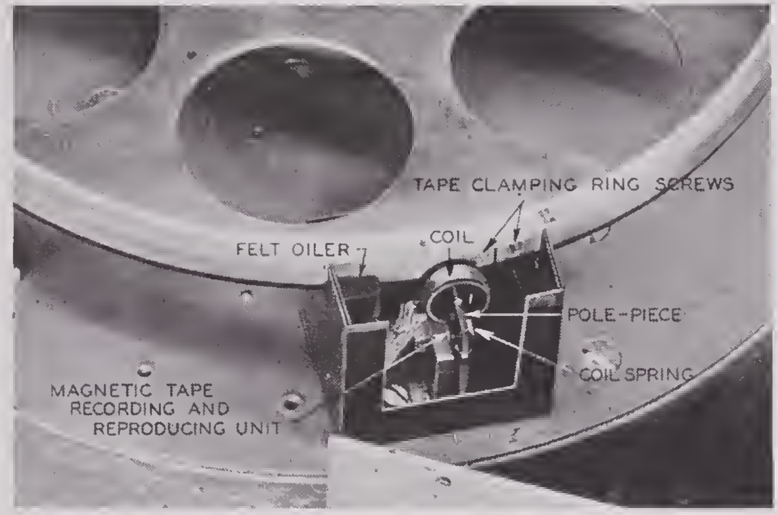

TAPE CLAMPING RING SCREWS
FELT OILER
COIL
POLE-PIECE
COIL SPRING
MAGNETIC TAPE RECORDING AND REPRODUCING UNIT

FIGURE 11. Close-up of magnetic tape recording and reproducing unit.

expected. The correct adjustment is one at which the transients, which are still present to some extent, will not be high enough in level to pass through the rectifier used to simulate the radio link. Thus the rectified output will equal the true pulse length. If the transient is added to the true rectified pulse, considerable time distortion results and may cause obliteration of a space. The same considerations apply to the adjustment of input level to the radio transmitter keying equipment. Since the level off the tape is substantially constant, a correct gain adjustment, once made, may be expected to remain correct.

21.5 APPARATUS DETAILS

A photograph of the flasher unit complete is shown in Figure 9. A view with the cover removed showing the arrangement of parts is shown in Figure 10.

21.5.1 Magnetic Tape Unit

One unit performs the function of recording, reproducing and erasing. This unit is shown in Figure 11. It consists of two pole pieces which bear on opposite sides of the magnetic tape. The coil is mounted over one pole piece. To obtain uniform magnetization of the tape at various frequencies, constant current should be supplied to the coil. When used as a reproducer, the coil must be connected to an impedance which will remain higher than its own impedance over the frequency range of interest in order to obtain substantially open-circuit voltage.

21.5.2 Tape-Recording System

The tape-recording system is actually composed of two independent recording circuits, (1) a low-speed recording circuit used to record the outgoing message on the tape, and (2) a high-speed recording circuit used to record the incoming message on the tape. These circuits are best discussed separately with reference to the schematic, Figure 12.

The frequency recorded at low speed is 60 cycles, supplied by one-half the 6.3-volt winding of the power transformer. This voltage is only applied when the flasher is in the low-speed dial positions LS and LR. It is also necessary that the low-speed timing cam contact T_7 be in a closed condition. The 60-cycle tone is keyed by the telegraph operator by means of the telegraph key. A pre-equalizing network C_3, L_1, R_2 is used to improve the frequency response characteristic of the tape unit. From the equalizer the keyed voltage is applied to the tape unit through switch S_{51} which must be on position 1 (dial position LS). It is well to note at this point that the 60-cycle tone is applied to the magnetic-tape unit only on dial position LS. On position LR the 60-cycle tone is applied only for the purpose of operating the low-speed timing light.

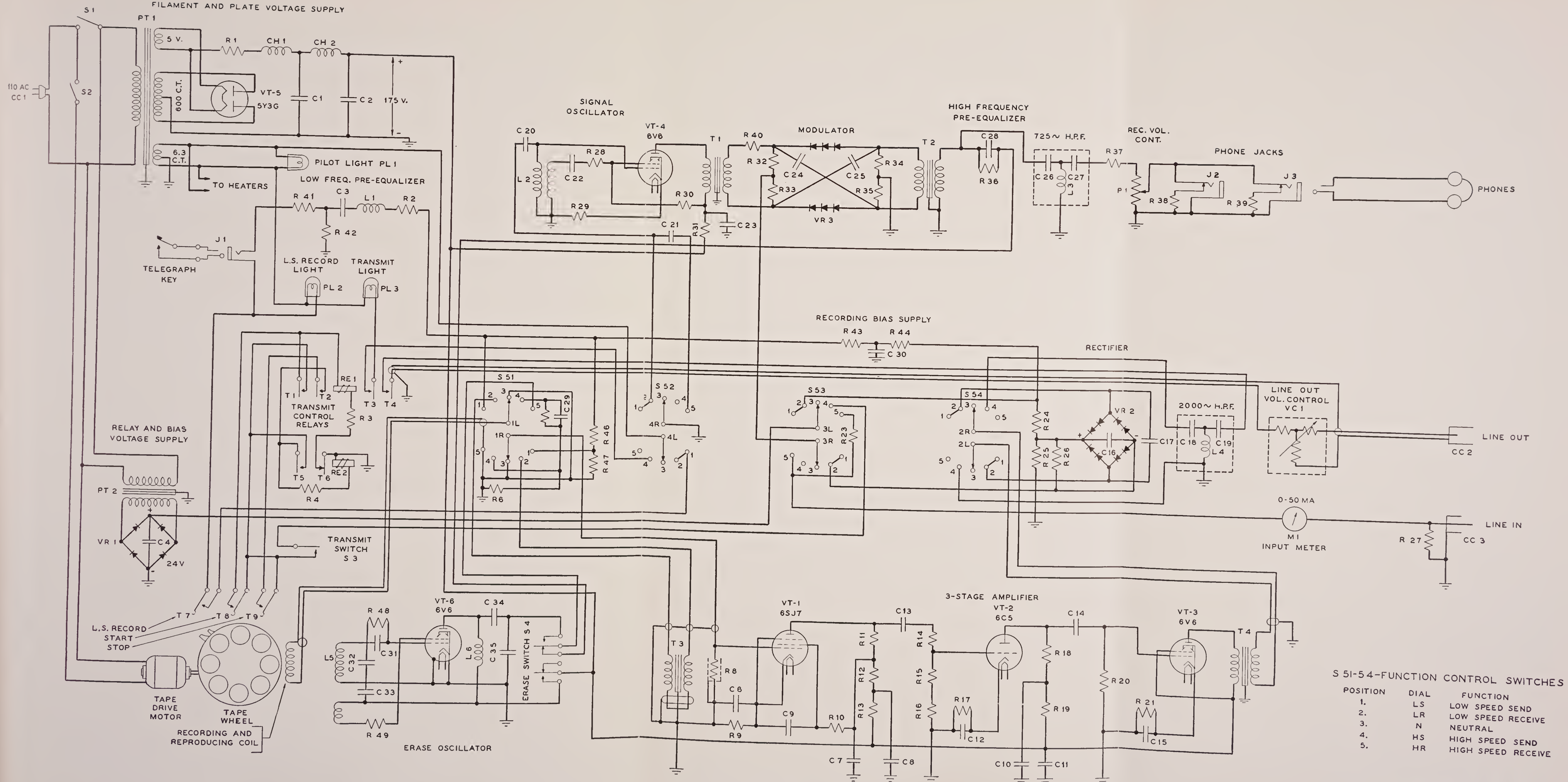
A d-c bias is provided for low-speed recording by the relay and bias-voltage supply PT_2 and VR_1. When the flasher is on dial position LS the rectified d-c voltage is applied through switch S_{53} to the recording bias-supply circuit, which adjusts the bias current to 1 ma and filters the ripple due to rectification.

Monitoring during the process of low-speed recording is provided for. Part of the recording voltage is applied through the bleeder R_{46}, R_{47} to the amplifier input.

At high speed the signal to be recorded consists of the pulses appearing on the incoming line. When the flasher is on dial position HR, these pulses are applied through switch S_{53} to the control circuit of the modulating or keying varistor VR_3, thereby keying the 6,000-cycle output of the signal oscillator. This is of the electron-coupled type using a 6V6 tube (VT-4). The keyed 6,000-cycle pulses pass through the pre-equalizing network R_{36} C_{28} to switch S_{51}, which applies the signal to the tape unit when the dial is on position HR. No recording bias is applied to the tape unit during high-speed recording. The receiving operator can monitor the incoming pulses by listening on the phones at the output of the modulator.

21.5.3 Tape Reproducing System

In picking up the signal which has been magnetically recorded on the tape, the voltage level generated is directly proportional to the tape speed. Thus in



S 51-54-FUNCTION CONTROL SWITCHES

POSITION	DIAL	FUNCTION
1.	LS	LOW SPEED SEND
2.	LR	LOW SPEED RECEIVE
3.	N	NEUTRAL
4.	HS	HIGH SPEED SEND
5.	HR	HIGH SPEED RECEIVE

FIGURE 12. Flasher unit schematic.

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changing the speed by a factor of 100/1 the level is increased by 40 db. Therefore, an appropriate amplifier-gain change has been incorporated in switching between low and high speeds. This operation is done by switch S_{51}. An input transformer (T_3) is used in reproducing at low speed when the highest level obtained from the tape unit is approximately 70 db below 1 volt. At high speed the unit is connected to a pad, which also serves to further equalize the high-frequency response, and thence directly to the grid of *VT*-1. The amplifier consists of three stages having a voltage gain of approximately 100 db in the high-gain condition and 55 db in the low-gain.

In the low-speed condition used in receiving a message on the headphones, the output of the amplifier is rectified by the bridge-type rectifier VR_2. This rectified voltage controls the modulator VR_3 passing the tone from the signal oscillator to the headphones. The operation of the modulator is briefly as follows. Each arm of VR_3 is a stack-up of copper oxide disks having unilateral conductivity. A bias voltage developed across R_{25} is supplied in the nonconducting direction, making the transmission path between T_1 and T_2 a high resistance. The rectified signal voltages are applied in the conducting direction, making the resistance low. Thus VR_3 acts as a variable attenuator whose loss is governed by the signal voltage opposing the fixed bias. The change of level obtained is in excess of 40 db. The frequency of the tone generated by the signal oscillator is approximately 1,200 cycles. A simple high-pass filter is included in the signal circuit to discriminate against 60-cycle ripple introduced in the modulator circuit. Volume control is provided by P_1.

At high speed the output of the amplifier is switched through a 2,000-cycle high-pass filter to discriminate against low-frequency noise. The signal frequencies of interest are 2,250 to 9,750 cycles. The line then goes through the timing relay to the line volume-control attenuator VC_1 and thence through the output line to the radio-transmitter keying circuit.

21.6 REQUIREMENTS ON RADIO CIRCUITS

Certain requirements are imposed on the radio circuits to employ successfully flash telegraphy. Most of these have been discussed previously, but because of their importance they are summarized here.

1. The radio transmitter must be equipped with electronic or other keying means capable of operation at the high keying speeds required by the flasher.
2. The tuning of the transmitter and antenna must be sufficiently broad to radiate side-band energy over a 7,500-cycle band, 3,750 cycles each side of the carrier.
3. The receiver must have broad enough tuning to receive a band width of from 5,000 to 7,500 cycles. Excellent low-frequency response is essential. For this reason it is desirable to take the output from the detector rather than after one or more stages of audio-frequency amplification.
4. Signal-to-noise ratio requirements are such that a signal having an average value of 6 db greater than the average noise level over a receiving time interval (0.4 second) will give satisfactory results. The instantaneous noise level during this period should not be greatly in excess of the average value.
5. Amplitude-fading and path-difference requirements are such that fading of less than 30 to 35 db and path differences of less than 50 to 100 μsec will result in satisfactory reception.

21.7 SUMMARY

1. Except in minor respects, the magnetic-tape flasher units discussed herein afford practical terminal equipment for achieving flash telegraphy at the high rate of 3,000 words per minute. If still higher speeds are of interest, it is believed that speeds up to 9,000 words per minute may be achieved by recording and reproducing d-c rather than a-c pulses, as is now done.
2. Recording at high speed without the use of a bias current permits a received flash message to be retained until erased by a manual operation and obviates the need, from a recording standpoint, of a starting or timing pulse.
3. The employment of an expanding limiting type of booster amplifier and associated modulator for receiving flash messages, provides considerable discrimination against static, multiple-path transmission, and fading.
4. To make flash telegraphy a dependable means of radio communication, affording good security against direction-finding determinations and interception, it is desirable to engineer a complete terminal unit, a radio transmitter-receiver system employing suitable radio frequencies, types of directive antennas, types of limiting amplification, and operating techniques. The dependability may be further increased by designing radio transmitters to flash large powers for short time intervals.

Chapter 22

FREQUENCY-STABILIZED MASTER OSCILLATOR

A study of instability as caused by the oscillator tube and the effect of circuit components; an investigation of the possibilities of multitube oscillators; development of the triode single-tube oscillator and buffer.

22.1 INTRODUCTION

At the time this project[a] was started, a tank transmitter required 125 crystals to cover all the possible frequencies to be used and, at the time, there was considerable doubt whether enough crystals could be produced in time for use in the war. It was thought that a small master oscillator could be designed to cover the frequencies from 2 to 20 mc with sufficient stability to meet the requirements of the Services.

Under the project, an oscillator followed by a buffer amplifier was developed which was small and had good frequency stability. This oscillator consisted of a Hartley circuit using a triode (6C4) followed by a 6AG7 buffer tuned to twice the oscillator frequency which delivered to a load consisting of 8,200 ohms and 20 $\mu\mu$f capacitance from 65 to 130 volts or 0.5 to 2.0 watts when powered from a 6- to 8-volt source. The stability is indicated by Table 1.

22.2 PRELIMINARY RESEARCH

Before development of an actual oscillator was undertaken, research was instituted into the share contributed to frequency instability by the tube itself as distinct from circuit components. This was followed by an investigation of the comparative advantages of one-tube oscillators and multitube circuits.

The tube investigation showed that frequency drift introduced by the tubes during the warm-up period was produced largely by inadequate and changing cathode emission. Other factors which contribute to frequency drift are mechanical expansion, distortion or buckling of the tube elements, and secondary emission from the bulb and mica parts.

22.2.1 Tube Studies

Effect of the tube on oscillator drift was studied in the following steps:

1. Self-ballasting filament. A tube with a single 0.008-in. carbon filament was constructed to investigate the usefulness of the negative temperature of carbon in combination with external resistance to provide ballast action. The effect was too small to be useful.
2. Anode expansion tubes with anodes of molybdenum, nickel, and copper to explore the effect of expansion on grid-plate capacitance. The low plate dissipation caused so little heating that the differences among the three sets of tubes was inappreciable.
3. Secondary emission. Secondary emission, probably from bulb and mica insulators, caused a warbling frequency shift. Carbonizing the bulb helped to reduce the frequency change, but lowering the anode voltage to 100 or 125 or lower was recommended for master oscillator tubes where frequency drift was to be kept to a minimum.

Table 1. Summary of stability tests.

Test	Per cent stability					
	2 megacycles		10 megacycles		20 megacycles	
	Avg	Max	Avg	Max	Avg	Max
Reset	0.0004	0.0011	0.00037	0.00065	0.00026	0.00045
Backlash	0.00019	0.00035	0.00017	0.00055	0.00039	0.00007
Tube change—oscillator	0.006	......	0.0175	......	0.02	......
Tube change—buffer	0.00068	......	0.0042	......	0.0069	......
Humidity	......	0.0019	......	0.0025	......	0.0069
Temperature	......	......	0.00078	0.0012	......	......
±5% line voltage						
1 minute	......	0.0065	......	0.0026	......	0.0037
5 minutes	......	0.0058	......	0.0044	......	0.0043
Drift						
5 minutes	......	0.0024	......	0.007	......	0.006
2 hours	......	0.0017	......	0.017	......	0.006

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[a] Project C-59, Contract OEMsr-690, RCA Laboratories.

4. Oscillator-coupling tube. A combined oscillator-buffer tube was studied but the decision to investigate possible improvement of the 6C4 was made rather than to pursue the combination tube further.

5. Mica-spacer slippage. Sudden discontinuous changes in frequency were traced to slippage of mica insulators. Polishing or reaming the holes in the mica helped the situation.

6. Advantage of generous emission. The warm-up time of a tube depends upon the cathode temperature. A heavier heater decreased the warm-up period but had no effect on the long-time frequency variations.

7. Thermal expansion of metal tube parts. Expansion of the cathode due to the high temperature at which it is operated changes the grid-cathode spacing and the capacitance between the two elements. Compensating capacitor disks attached directly to the cathode made it possible to produce a tube with practically zero temperature coefficient of capacitance with respect to heater voltage. Since the grid cathode is across few turns of the circuit inductance, a large improvement in stability is needed before the overall frequency drift can be improved much by this expedient.

No evidence was obtained for frequency change due to anode expansion. Grid expansion is more serious and some work was carried out to decrease frequency shift from this cause. Invar could not be employed because of the high temperature of the grid wires; molybdenum was finally recommended.

22.2.2 Conclusions of the Tube Research Group

Frequency drift in oscillators is partly chargeable to the tube and partly to the circuit in which the tube is used. The changes in frequency produced by electronic effects in the tube may nearly always be compensated by proper circuit design. These tube effects are well understood. Defects in the tubes which are subject to correction are mostly of a mechanical nature. Actual thermal expansion of the tube elements is not sufficiently great to cause serious capacitance changes in well-designed circuits, but if this expansion is magnified by mechanical leverages very large changes in tube capacitances may occur. Close spacing of tube elements makes the tube capacitances critically dependent upon maintenance of those spacings. Complex geometrical shapes of the elements produce unexpected and unpredictable distortions and should be avoided. For use in a stable oscillator, the tube should have ample emission so that the number of electrons in the interelectrode space will be dependent only on the electrode potentials, the effect of which on frequency is predictable and may be nullified by proper circuit design. Low plate voltages, 125 and below, are important to reduce secondary emission which has considerable effect upon frequency stability.

Frequency drift due to tubes alone, computed from 15 seconds after the filament voltage is applied to the oscillator, can be held well within 0.002 per cent.

22.2.3 Studies of the Oscillator Circuit

An experimental study was made to determine the limitations imposed on frequency stability by standard oscillators of the Colpitts and Hartley types, with particular interest in the warm-up period. Other investigations related to frequency versus filament and plate voltages, use of powdered-iron cores in the tuning coils, stability with output load coupling, and capacitance irregularities in ceramic and mica fixed-tuning capacitors.

To avoid frequency drift due to heating of components, the tube socket was mounted 3 in. above the circuit parts at the end of wire rods. In addition, ceramic sockets which did not cause appreciable drift in frequency with temperature change were used, but even with these precautions there was appreciable drift due to the heat generated in the circuits by oscillations. This was overcome by operating the circuit for an hour with a spare tube prior to taking data. No attempt was made to temperature-compensate the components. Ceramic capacitors with approximately zero temperature coefficient were used. Tuning was accomplished by moving a powdered-iron core within the coil by a micrometer adjustment. The choke coil for plate feed consisted of three universal-wound sections having low distributed capacitance and a natural period of approximately 1,400 kc.

The data in Figure 1 are characteristic of a Colpitts tapped-capacitance type circuit. The Q of the circuit was about 140 and the total capacitance of the resonant circuit was about 270 $\mu\mu$f. The impedance of the tuned circuit was about 40,000 ohms. Figure 2 gives the same data at 11 mc. In this case the inductance was adjusted to resonate with the same 270 $\mu\mu$f of capacitance employed in the 2-mc tests. Circuit Q was about 200, impedance about 10,800 ohms.

Similar studies using the Hartley circuit gave about the same results as with the Colpitts circuit.

In each case use of a 955 tube gave somewhat greater stability than the 6C4.

22.2.4 Component Effects

Some of the ceramic capacitors were quite unstable even though of the same manufacture and same general type. Others were quite stable. Cores of copper, magnetite, powdered iron, and a combination of copper and magnetite were interchanged and the effects upon frequency change noted. It was discovered that a proper combination of copper and magnetite could be found which would give practically any degree of compensation with regard to changing plate voltage that might be desired.

Over a test period of 120 hours, no effects due to permeability-aging of either powdered-iron or powdered-magnetite cores were noted.

22.2.5 Oscillator plus Buffer Stage Stability

The curves in Figure 3 give a general idea of the frequency drift of oscillator and output buffer stage. In this case the oscillator was loosely coupled to the 6AG7 buffer acting as a class A amplifier. With 125 volts on the oscillator plate, 1.25 volts were fed to the buffer grid circuit and 0.6 watt was developed in the buffer output load of 7,100 ohms. With 44 volts to the oscillator plate, the oscillator delivered 0.28 volt to the buffer which developed 0.035 watt in the load.

22.2.6 Conclusions of the Circuit Studies

Frequency instability contributed by the tube in an oscillating circuit is caused mainly by changes in tube input and output resistances produced by changes in plate voltage or cathode emission. In this manner a phase shift of the regenerated oscillations occurs. This calls for an opposite phase shift in the resonant circuit which is produced by operating at a different frequency.

Proper choice of circuit elements can compensate frequency changes due to tube changes. Circuit Q should be as high as practical. Small tubes like the

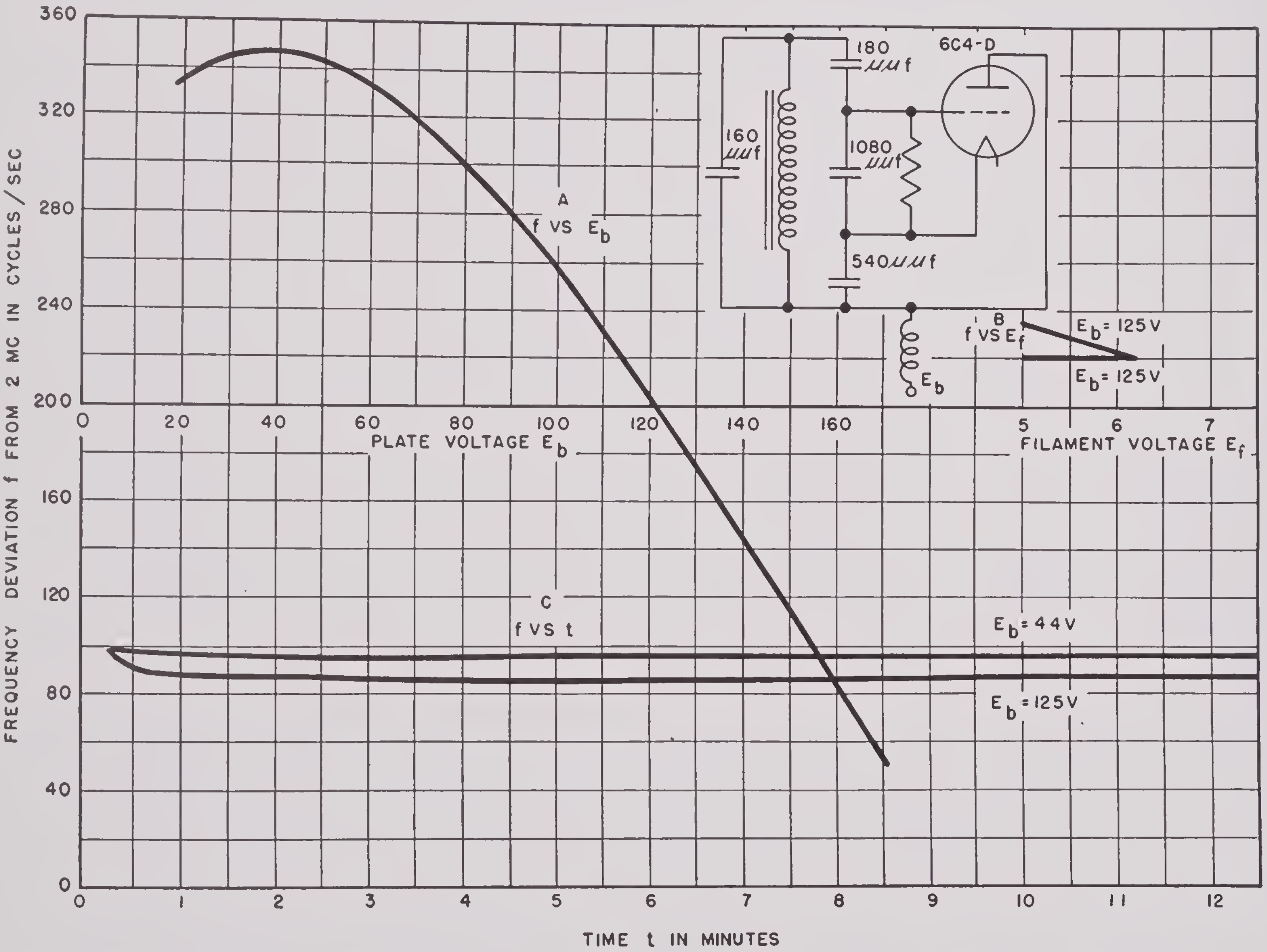

Figure 1. Experimental data showing effect at 2 mc of plate and filament voltages on frequency and magnitude of frequency drift.

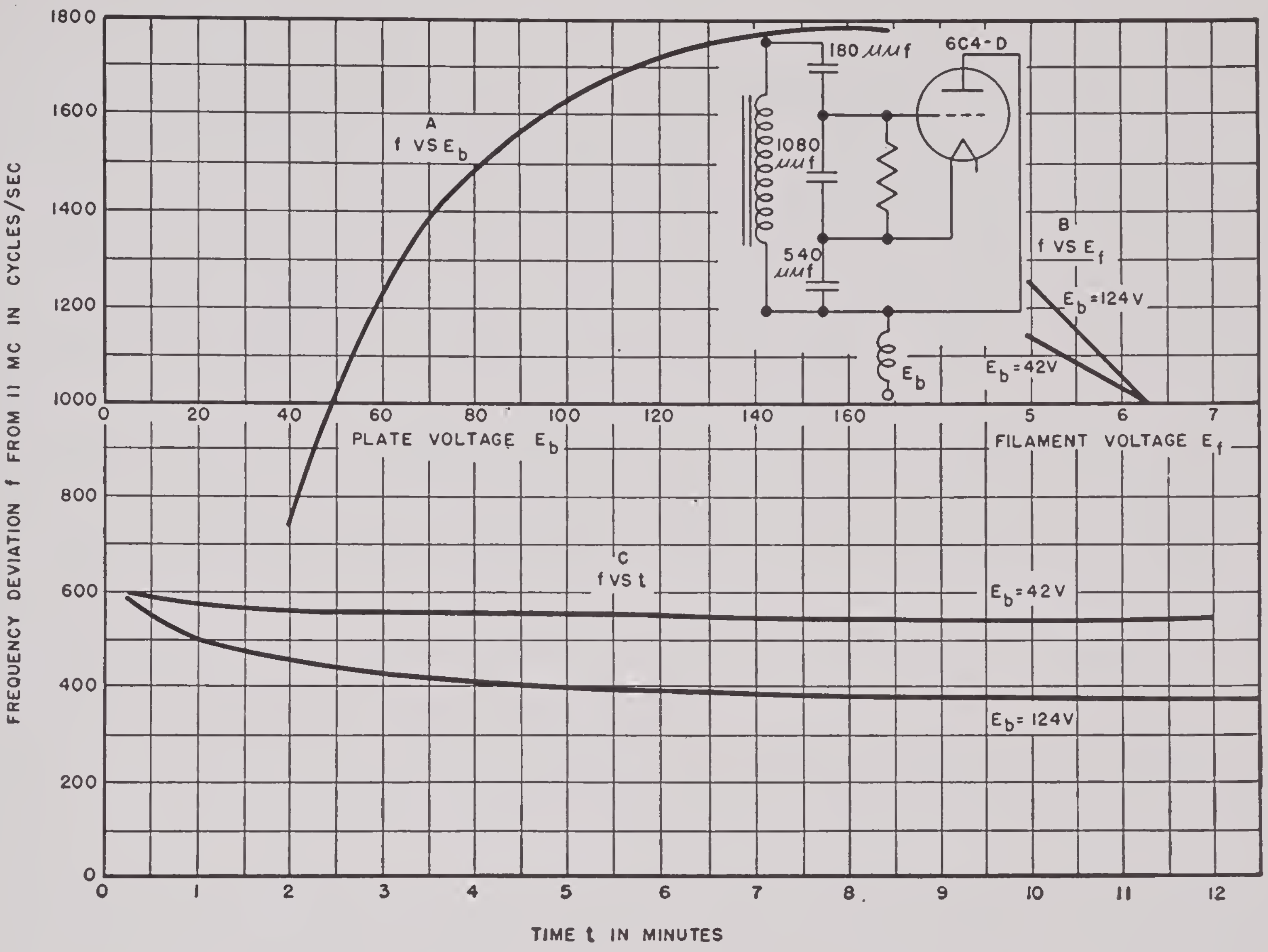

Figure 2. Frequency stability with respect to time and plate and filament voltages at a frequency of 11 mc.

955 or the 6C4 should be used. The tube should be tapped across portions of the tuned circuit which are as small as is consistent with stable operation. A grid-circuit load of 500 ohms and a plate-circuit load of 2,000 ohms seem to be a satisfactory arrangement. Low plate voltage is desirable; grid-leak resistance value is not critical. Coupling to the output stage should be low. Operating the buffer on the same frequency as the oscillator, which had a plate voltage near the upper limits used, made it possible to deliver a maximum of about 1 volt to the buffer.

The change in permeability of a powdered-iron core in the tuning coil with change in r-f magnetic field can be utilized to improve the frequency stability with changes in plate voltage.

22.3 ADDITIONAL PRELIMINARY STUDIES

Several multitube oscillator circuits were examined, but the conclusion was reached that better frequency stability could be achieved with well-designed single-tube oscillators.

22.3.1 Oscillator Plus Amplifier

For example, looser coupling to the resonant circuit can be employed if the transconductance of the oscillator is higher. Therefore, if the oscillator could have its gain supplemented by an amplifier, very loose coupling could be used. But if there are other circuit phase shifts, the virtues of this system would not be realized.

22.3.2 Low-Frequency Amplifier Plus Heterodyne Detectors

Amplification with minimum phase shift can be accomplished by using a low-frequency amplifier. Two heterodyne detectors are required to accomplish the two necessary frequency conversions. In a case examined, two i-f stages of high L/C ratio were employed, each stage having a single tuned circuit. One stage was tuned below the nominal intermediate frequency (0.29 mc) and the other was tuned above the i-f frequency (0.39 mc). The local oscillator frequency was

9.32 mc and the stabilized output was 9 mc. At the mean frequency of the amplifier the coupling circuits acted like nearly pure reactances so that a change in transconductance of the tube or in tube capacitance or a change in intermediate frequency should produce very little change in phase shift.

In this case, a change in local-oscillator frequency of 28 mc produced a change in output frequency of only 2 mc.

22.3.3 Double-Heterodyne Oscillator with AFC and AVC

Adding automatic frequency control to the circuit described above decreased appreciably the frequency shift. A maximum change of approximately 400 cycles from 10 mc was produced by plate voltage changes up to 30 per cent.

22.3.4 Oscillator with AVC

Application of automatic volume control to a Colpitts oscillator to prevent grid-current flow did not improve the frequency stability, indicating that, in the usual oscillator circuit, a better balance of input capacitance changes and better frequency stability are obtained when the grid is permitted to go slightly positive every cycle.

22.3.5 Bridge-Stabilized Oscillator

Another method to improve stability was by means of a thermally controlled resistance element which affected the balance of a bridged-T network in accordance with the amplitude of oscillation. Instead of the quartz crystal element originally employed by Meacham,[1] customary circuit elements were used. This circuit was more nearly independent of plate-voltage changes, although frequency change occurred as with the other multitube circuits if the filament voltage changed. Long time variations were much less with the bridge-stabilized oscillator, 10 to 20 cycles' change being noted after a period of about 1 hour.

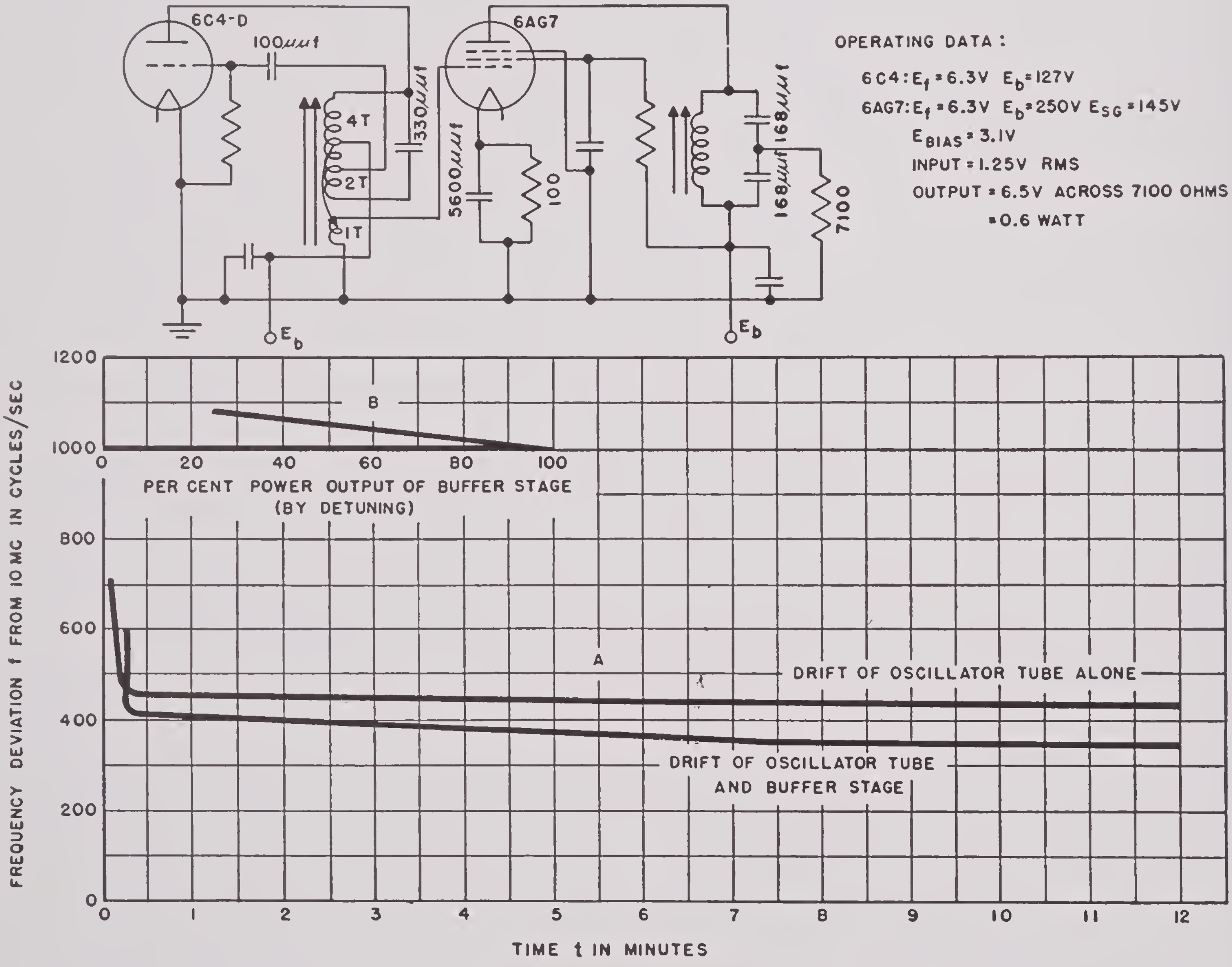
6C4-D
100μμf
4T
2T
1T
330μμf
6AG7
5600μμf
100
168μμf
168μμf
7100
E_b
E_b
OPERATING DATA:
6C4: E_f = 6.3V E_b = 127V
6AG7: E_f = 6.3V E_b = 250V E_{SG} = 145V
E_{BIAS} = 3.1V
INPUT = 1.25V RMS
OUTPUT = 6.5V ACROSS 7100 OHMS
= 0.6 WATT
1200
1000
800
600
400
200
0
B
0
20
40
60
80
100
PER CENT POWER OUTPUT OF BUFFER STAGE (BY DETUNING)
FREQUENCY DEVIATION f FROM 10 MC IN CYCLES/SEC
A
DRIFT OF OSCILLATOR TUBE ALONE
DRIFT OF OSCILLATOR TUBE AND BUFFER STAGE
0
1
2
3
4
5
6
7
8
9
10
11
12
TIME t IN MINUTES

FIGURE 3. Characteristics of an oscillator buffer stage operating at 10 mc.

22.4 DEVELOPMENT OF THE C-59 OSCILLATOR

Since changes in the load circuit of any oscillator must have minimum effect on the frequency generated if high-frequency stability is desired, the first decision in developing an actual working model of a highly stable oscillator was to use a buffer amplifier tuned to twice the frequency of the oscillator. In this manner, the effect of the load upon the oscillator can be reduced by a factor of 100 to 1 for the reason that the output circuit of the buffer looks to the oscillator like a very low value of inductive reactance and variations in this reactance, which are low, have very little effect upon the oscillator frequency.

22.4.1 Temperature Effects

Temperature changes play an important part in instability of oscillator frequency. There are two separate causes, change in temperature of a tube or component due to the self-heat generated by a current flowing through it and change in ambient temperature. Both conductors and insulators are subject to this effect, the first because they have resistance and the second because of dielectric hysteresis or power factor. Power factor is practically independent of capacitance, voltage, or frequency but is influenced by temperature, nearly always becoming higher as the temperature is raised.

The temperature coefficient of an oscillator tank circuit built up of commonly used parts will fall in the range of 10 to 100 parts per million per degree centigrade (ppm/°C). Variations in resistors, other dielectrics in the circuit and the tube will bring this value to the range of 10 to 500 ppm/°C. Some of the elements will have positive and others negative coefficients. While there is no simple way of reducing the temperature coefficient of inductors, capacitors are available having a wide range of temperature coefficients ranging from -750×10^{-6} to $+120 \times 10^{-6}$ $\mu\mu f/\mu\mu f/°C$.

Most of the effect of temperature tends to lower the frequency of an oscillator so that capacitors having negative temperature coefficients can be used to correct this effect to some extent, particularly when ambient temperature changes are to be compensated. The use of such capacitors produces warm-up drift, however.

Resistors of the metallized-filament type have low temperature coefficients and are useful in the present problem.

22.4.2 Effect of Humidity

Changes in humidity produce frequency changes because of (1) change in the dielectric constant of air, (2) change in the surface resistivity of solids, and (3) change in the volume resistivity of solids.

The first factor varies from approximately 1.0005 to 1.0016 over a humidity range from 0 to 100 per cent and a temperature range of from 0 to 60 C.[8]

Since both L and C are affected by humidity, only two solutions are possible. Either make L and C so that humidity does not affect their constants or isolate humidity so that it does not reach L and C.

In a typical 10-mc oscillator the frequency varied over a range of 0.002 and 0.05 per cent when the humidity was varied from 30 to 100 per cent. This oscillator was substantially airtight. Gaskets were used on all covers and packing was used around the control shafts. Entering leads were sealed with glyptol cement. All coil forms and solid dielectrics were of ceramic or mycalex. A dessicant of silica gel was placed in the interior of the oscillator.

22.4.3 Importance of High Q

Although the variation of voltage on filament and plate elements of the tube and the magnitude and constancy of the applied load and of other circuit elements have important bearings on the frequency constancy, the effective Q of the tank circuit is perhaps the greatest factor having to do with the generation of constant frequency. Frequency stability is directly proportional to the Q of the tank circuit, since the frequency change necessary to compensate the phase shift resulting from changes in load, tube resistance, and other circuit resistance is proportional to $1/Q$. Thus Q should be as high as possible.

22.4.4 Tuning Methods and Mechanisms

Since high values of Q are so important and since a miniature oscillator presents new problems of obtaining good Q with small coils, a study was made of several methods of tuning the tank circuit.

Rotary-coil tuning gives a large inductance ratio, uniform Q over the tuning range, but poor reset because of sliding contact between winding and slider. Variometer tuning gives a large inductance ratio but very poor Q. Flat-spiral coil with copper-disk tuning gives a low inductance ratio, poor Q and a nonuniform tuning curve. Copper-plug tuning gives a low inductance ratio, somewhat higher Q than with the methods mentioned just above, and a Q ratio which is approximately of the same magnitude as the inductance ratio. Its great advantage lies in the fact that the fineness of tuning and reset value are limited only by the mechanical perfection of the threaded actuating mechanism.

Since the oscillator was to cover a wide range (2 to 20 mc) and because of mechanical limitations in making one tuning mechanism to cover the entire range continuously, it was decided to break up the range into bands, assuming a frequency change of 0.02 per cent per dial division, dial divisions being limited to 1,000. Thirteen sets of coils were required for the oscillator and five for the buffer.

22.4.5 Final Oscillator Tuning Arrangement

The final design of the tank circuit for band 13 (17.50 to 20 mc) is shown in Figure 4 to consist of the inductance properly tapped and the tank tuning capacitance of the ceramic tubular type having, theoretically, zero temperature coefficient. Actually two such capacitors were connected in parallel and so chosen that the effective temperature coefficient of the combination was close to zero. The capacitors were rigidly mounted to the coil form platform by a metal end-on capacitor fastened with a machine screw to the platform.

The coil form was molded from Styramic, a chlorinated diphenyl-polystyrene compound having a higher heat-distortion temperature and better machinability than polystyrene. The conductor was cemented to the form with polystyrene cement. The coil was held in the oscillator by means of three ball feet with threaded adjustment for length which allowed the tank assembly to be moved ±⅛ inch from the nominal position with respect to the tuning plug. This adjustment provides a ready means of producing oscillators which will have physically identical coils and which will produce the same frequency at a given dial setting for each oscillator without the necessity of altering the coils.

The exact position of the taps for grid, cathode, and plate was determined for each band to give the smallest frequency change with line-voltage variations. The temperature coefficient of the combination of coil, mounting, and tuning plug mechanism was within ±6 ppm/°C in frequency. The coil alone had a coefficient of ±1.5 ppm/°C. This was attained by proportioning the length of the coil to its diameter so that the inductance was decreased due to lengthwise expansion at the same rate as the inductance was increased by radial expansion. For the combination of copper wire, close wound on a Styramic form this ratio was 1.75.

The ±6 ppm/°C spread was due to the fact that the overall expansion of the tuning plug mechanism was different when the plug was withdrawn than when it was inserted in the coil. This effect was minimized by supporting the copper tuning plug on a pillar of Styramic, which did not touch the lead screw except at a point where the length of the pillar was cor-

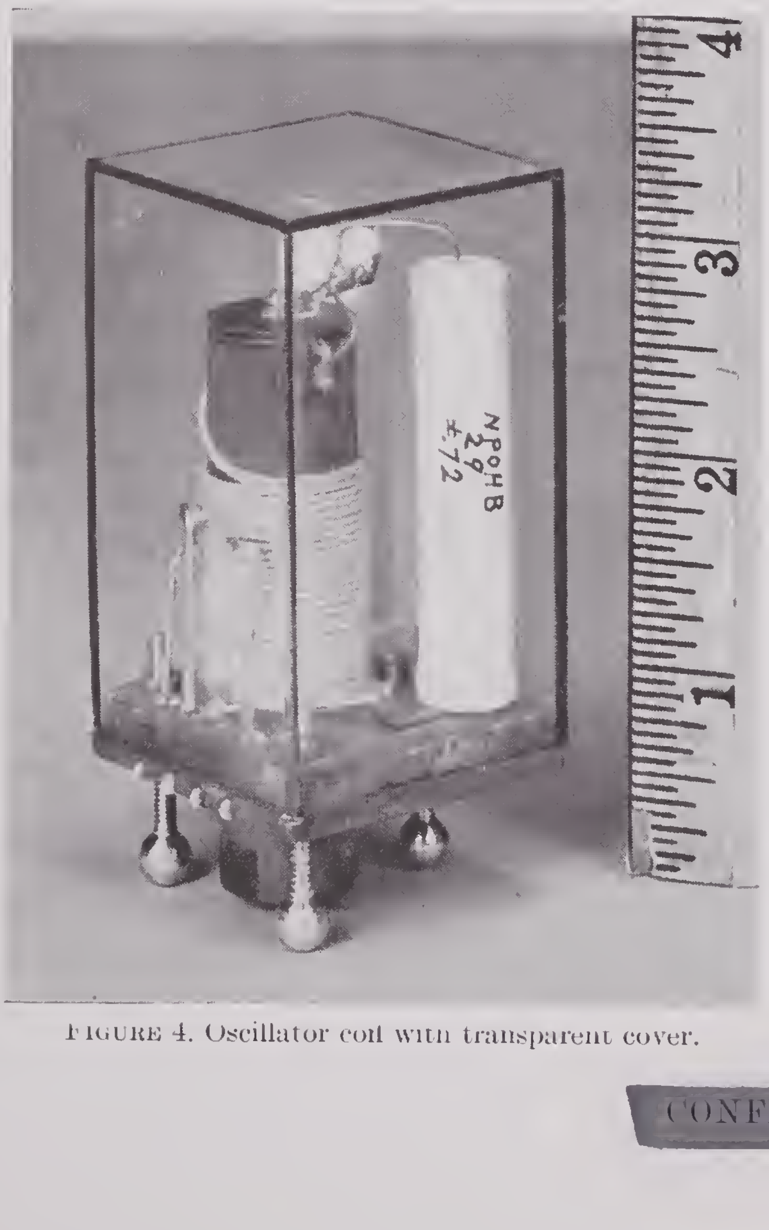

FIGURE 4. Oscillator coil with transparent cover.

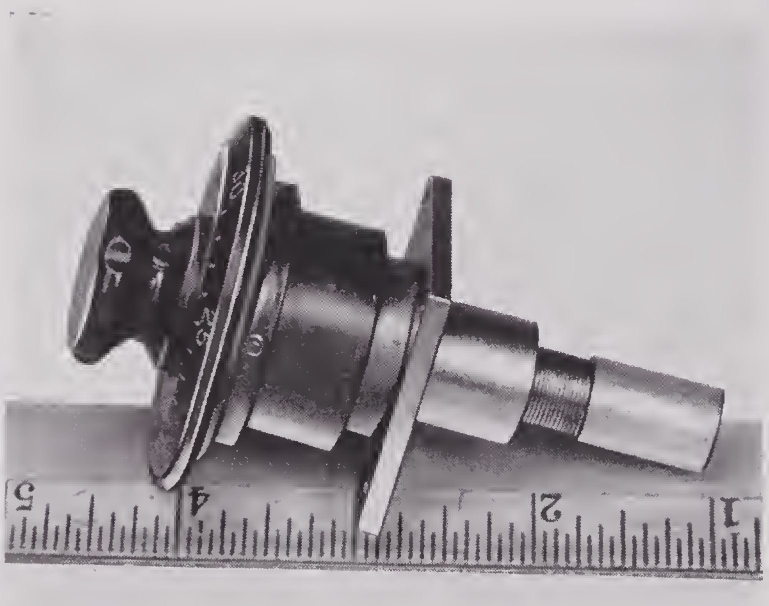

FIGURE 5. Tuning mechanism and dial, assembled.

rect to counteract the expansion of the coil from its ball feet to its electrical center, lengthwise.

Coil and capacitance were sealed in airtight containers.

To reduce further the effect of humidity on frequency, the oscillator-buffer unit was made as tight as possible without resorting to sealed joints. A container of silica gel was attached to one end of the chassis with provision for removing it so that it could be baked for 3 hours at 300 F to prepare the crystals for another cycle.

A great many measurements were made on this oscillator-buffer unit to determine its performance. The results of these measurements employing test specifications for checking Navy transmitters are summarized in Table 1.

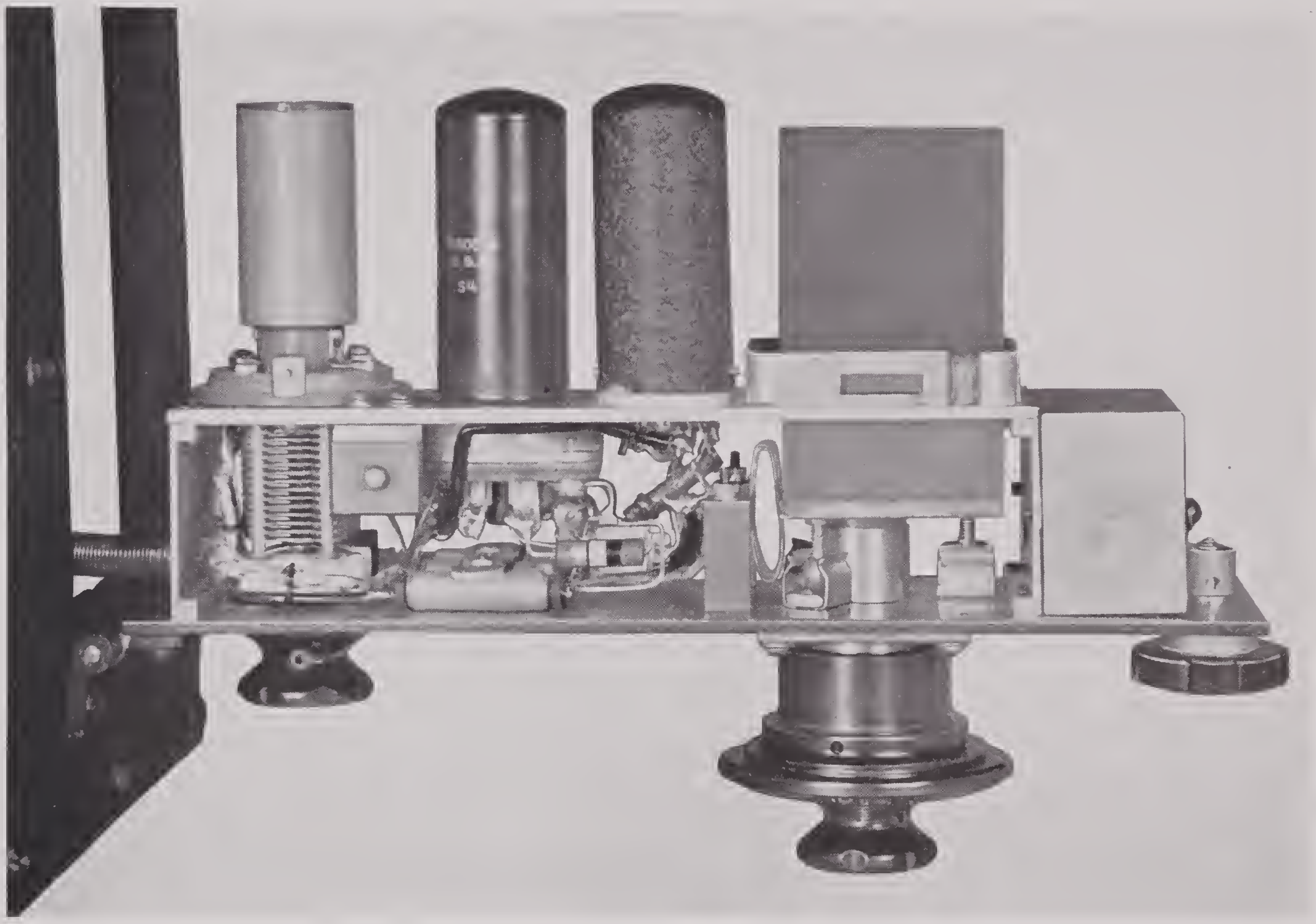

FIGURE 6. Oscillator, bottom side plate and shield, cans removed. Entire oscillator and buffer assembly requires 7⅛ x 2⅛ x 3⅞ in. of space and weighs 3 lb with 1 set of coils and tubes. Coils to cover entire range of 2 to 20 mc weigh 2.1 lb, and coil-carrying case without coils weighs 4 lb.

Chapter 23

PICKUP TUBE FOR RECONNAISSANCE TELEVISION

A line-mosaic iconoscope or television pickup tube for use in reconnaissance. Several tubes were constructed but the project ended with the feeling that to eliminate leakage between the elements would require considerable research and that an equal amount of effort would develop a two-dimension tube with greater sensitivity.

23.1 INTRODUCTION

PROJECT C-62[a] CALLED for the investigation of a line-mosaic pickup tube to determine whether or not it would be suitable for application in a high-definition (1,000-line) reconnaissance television system.

Such a system requires a higher definition than can be obtained from the normal iconoscope, the latter being primarily limited by the size of the scanning spot produced by the electron gun. Since it was felt that a rather extensive research would be required to increase the definition of this type of iconoscope or a low-velocity iconoscope to a point where it could be used for this purpose, an investigation of the possibilities of the line-mosaic tube might obviate this expenditure of time and facilities if it were found that a tube of this type could be used.

A line-mosaic pickup tube consists of an electron gun capable of producing a very narrow scanning spot and a mosaic made up of a row of fine vertical line elements. A signal plate on the back of the mosaic is connected to the external signal lead and serves as capacitance coupling to the elements. The scanning spot is deflected across the mosaic by means of a magnetic deflecting yoke.

The optical system used with this pickup tube is different from that of a conventional iconoscope, inasmuch as it must provide vertical deflection. This can be done as follows: An objective lens images the scene to be transmitted onto a slit image plane which permits the light from a single horizontal line of the image to reach the mosaic. The image as a whole is moved vertically across the slit at frame frequency by a rocking mirror located between the objective and the image plane. To simplify the geometric arrangement of the optical system, a weak cylindrical lens may be used in combination with the objective so that picture elements are imaged as horizontal lines on the surface containing the slit and as vertical lines on the mosaic.

The frame frequency used is much lower than is used in commercial systems. Its maximum is such that the picture can be transmitted over a channel having normal band width and the frame frequency may be much lower, depending upon the sensitivity needed, the means of reconstructing the picture and other factors. In view of the slow repetition rate, the picture cannot be reproduced on an ordinary kinoscope, but instead it is formed on a tube having a long decay period, or photographically, or by mechanical recording.

A pickup tube based on a line mosaic obviously cannot employ surface storage of the photoelectric charge. Instead the charge is stored for the duration of a single line. This reduces the intrinsic sensitivity of the pickup device. However, since the design of the tube is such as to permit saturated photoemission and because the frame frequency is low, the overall sensitivity can be made comparable with that of the normal iconoscope.

Prior to the start of the investigation, slit-aperture guns (electron guns having a fine rectangular slit at the crossover) had been tested and found to have a horizontal resolution of well over 1,000 lines. Therefore, this type of gun could be used without change in the line-mosaic pickup tube. Furthermore, a wash-off relief photographic process had been developed which, with slight modification and the working out of a specific technique, seemed to offer a method by which the fine line structure of the mosaic elements could be readily made.

23.2 LINE-MOSAIC PICKUP TUBE

The arrangement of the various elements making up the line-mosaic pickup tube can be seen in Figure 1.

As has been stated above, a slit-aperture gun has sufficient resolution in the horizontal direction to meet the requirements of the line-mosaic pickup tube. This gun employs a two-lens electron optical system, the cathode, grid, and first anode forming one lens, and the field between the first and second anode, the other. Electrons from the cathode are converged into a crossover near the first lens, this crossover being the exit pupil of the first lens system and the narrow-

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[a] Project C-62, Contract OEMsr-706, RCA Laboratories.

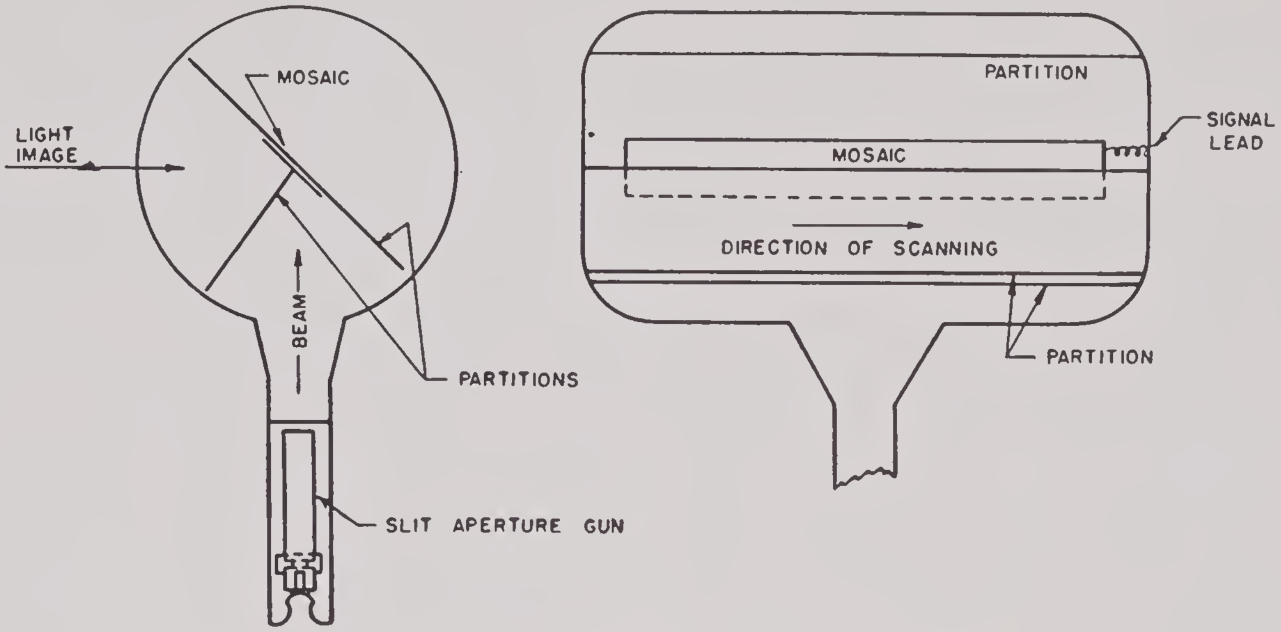

Figure 1. Elements making up the line-mosaic iconoscope.

est portion of the beam on the cathode side of the second lens. A diaphragm having in it a rectangular aperture approximately 5 mils high and ½ to 1 mil wide is placed at the crossover. The second lens is adjusted to image this aperture on the mosaic, so that the resultant spot is a short narrow vertical line. It should be pointed out that the spot is larger than the true geometric image of the slit aperture, because of the spherical aberration of the second lens, and consequently, if the slit size is reduced, the current reaching the spot is reduced without appreciable reduction in size, and for this reason, the redesign of the electron gun to give a thousand-line resolution in both directions represents a difficult research problem.

The scanning spot is deflected horizontally by means of a conventional magnetic deflecting yoke and suitable driving circuits. No new techniques are involved in the horizontal scanning operation.

23.2.1 Mosaic Types Investigated

Two types of mosaic were tried in the investigation. Both used a glass plate 30 mils thick as dielectric but differed in the arrangement of elements. One of them consisted of a row of fine line elements insulated from each other. The elements were about ¼ in. in length with 600 to the linear inch. Platinum was used as the base of the line elements, which were photosensitized with silver and cesium during the processing of the tube. A horizontal strip of silver on the back of the glass dielectric serves as the signal plate. The arrangement of this mosaic is illustrated in Figure 2A.

The second type of mosaic employed a barrier grid to prevent electrons scattered by one element from reaching the others. The nature of the barrier grid can be seen from Figure 2B. As before, the mosaic consists of line elements, but between each pair of elements is a conducting strip of connected metalized bands at either end. These shielding strips not only serve as barrier grid but also as the signal plate.

The elements of either type of mosaic are returned to their normal or reference potential each time the

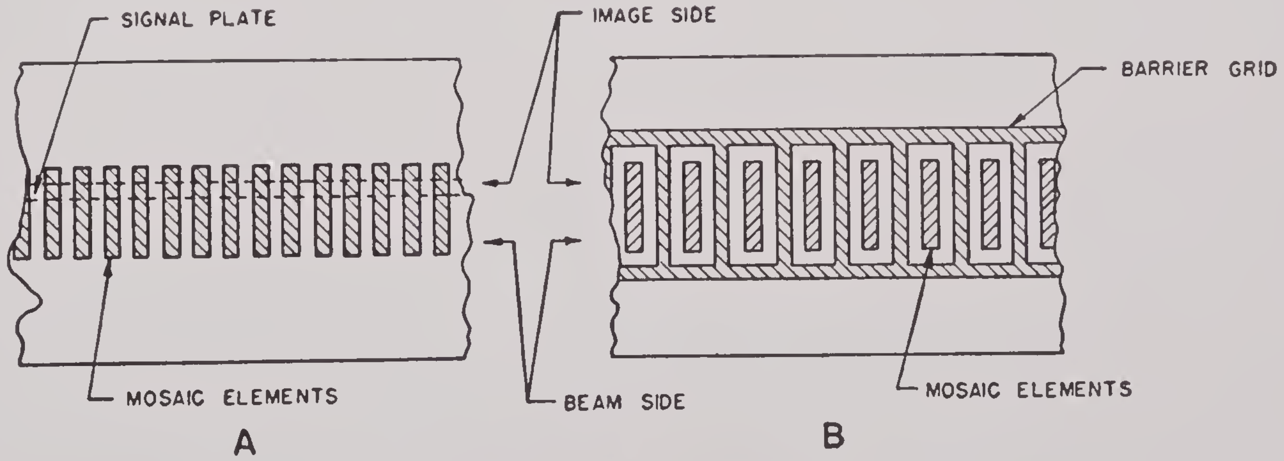

Figure 2. Details of signal plate for line-mosaic tube.

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beam passes over them by an equilibrium between the secondary emission electrons collected and the beam electrons. This means that the element potentials will be approximately that of the secondary electron collectors and that these collectors cannot be used to obtain saturated photoelectric emission. Therefore, the tube is divided into two compartments by a partition at right angles to the lines forming the elements. The beam strikes only the portion of the elements extending into the lower compartment and this compartment contains the secondary emission collector (i.e., the second anode). The upper compartment contains the photoelectric collector, and the light image is projected on the portion of the lines extending into this compartment. The photoelectric collector can be made positive with respect to the secondary-emission collector without disturbing the secondary-emission equilibrium of the mosaic elements. Therefore, the photoelectric emission from the illuminated elements can be saturated. Reference to Figure 1 will make clear the arrangement of partitions and compartments.

23.2.2 Construction of Pickup Tube

Two line-mosaic tubes were built to test the feasibility of the pickup devices described above. Each tube contained sections of ordinary and barrier-grid line mosaics, thus making it possible to test the two types under the same conditions of activation and operation.

The construction of the gun and glass envelope followed conventional lines and will not be further described.

The preparation of the mosaic involved a special photographic washoff relief procedure which was in part developed for the purpose. A plate of 30-mil glass, 4½ in. long and 2¾ in. wide formed the base of the mosaic. After being thoroughly cleaned, the glass was coated with a sensitizing solution prepared as follows:

Gum-Bichromate Process Solution

Solution I 30 g gum arabic in 100 cc water
Solution II 5 per cent potassium dichromate

One part of Solution II is added to two parts of Solution I. The formula produces a gum-arabic film of low sensitivity but having a very high contrast and resolution, which is required for the line structure.

The solution is flowed over the glass plate at room temperature and let to drain until dry. When dry, the plate is found to be covered with a thin transparent film of sensitive material which hardens when exposed to the light of the short-wave end of the visible spectrum.

The negative used for preparing the line structure was a contact print transparency made from photographically reproduced copies of a 600-line ruled grating. The negative had 600 lines per inch, covering an area 4 in. long by ¼ in. high.

The sensitized plate was exposed to the light of a mercury arc through the 2,400-line negative and then was developed in water at 10 C until the unexposed portions of the sensitive film were completely washed away. Platinum was then sputtered on to the relief image thus obtained until the transmission was reduced to 10 per cent. The plate was then immersed in boiling water. This dissolved the gum-arabic lines in the exposed portions and carried away the platinum covering them. An array of sharply resolved platinum lines suitable for the mosaic remained on the glass plate.

On the half of the mosaic to be used as a barrier grid, a platinum strip, in the form of reduced Hanovia Platinum Bright painted on the glass, is made so that it is just in contact with the lines. Every other line is then cut mechanically with the aid of a fine needle so that it does not make contact with the platinum strip. These electrically insulated lines serve as the mosaic elements, while the lines remaining in contact with the strip form the barrier grid. A photographic process could be used giving alternate long and short lines if it was desired to produce these barrier-grid mosaics in any quantity.

On the back of the glass, opposite the portion of the mosaic which did not have the barrier grid, a signal plate was made in the form of a strip of reduced platinizing solution.

The first mosaic to be formed in this way was then placed in an evaporating chamber and a very thin layer of silver was deposited on the surface. This silver film was so thin that its electrical conductivity was not measurable by ordinary methods. The mosaic was then sealed in a glass envelope with a gun, collector electrodes, and partitions arranged as described above. During the preliminary exhaust the tube was baked at 450 C for more than an hour. This removes contamination and gas from the glass and metal parts and also will further reduce the conductivity of the silver film by breaking it up. The activation procedure which was followed was the usual oxidation of the silver, addition of cesium vapor, and heat treatment.

Since the sensitivity of this tube was very low and there was considerable evidence of leakage between elements, a second tube was built. The platinum line structure was formed in the same manner but silver was not deposited on the platinum outside the tube. Instead, silver evaporators were mounted inside the glass blank in such a way that silver could be deposited after the initial exhaust was complete. The silver layer used was considerably thinner than that in the previous tube. Activation was carried out in the same way as before. The sensitivity of this tube was considerably higher than that of the first tube.

Figure 3 is a photograph of one of the line-mosaic pickup tubes.

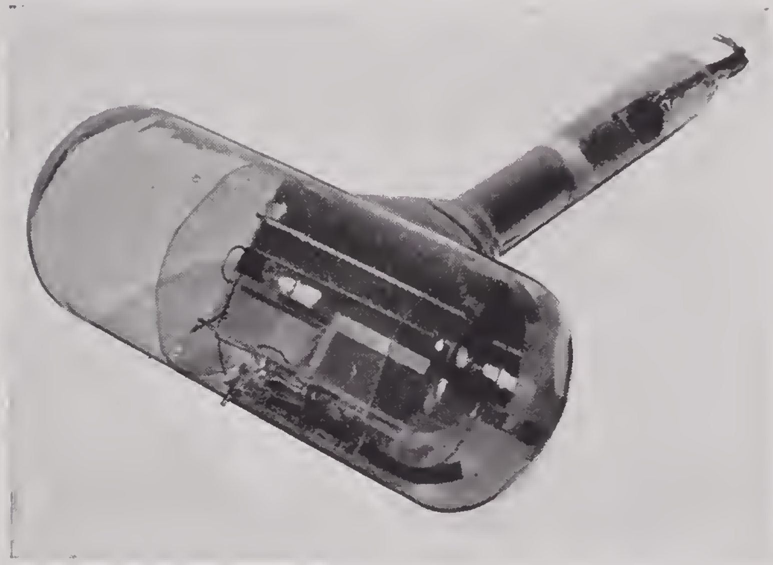

FIGURE 3. Photograph of one of reconnaissance television tubes.

23.2.3 Experimental Results

The equipment used to test the performance of the line-mosaic tubes operated with a horizontal deflection frequency of 15 kc, but since the amplifier response extended to a high frequency and the amplitude could be decreased at will, the test equipment did not limit the ultimate resolution of the system.

The first tests were to determine whether or not the gun was capable of resolving the line structure. It was found that the line structure could be resolved, but that the actual beam size was somewhat larger than the element width. Since an actual mosaic of this type would have two or three line elements per picture element, the spot size was considered adequate.

Tests were made with a line of light of variable width to determine the definition in terms of the sharpness of the edge of the reproduced image of this line. Overall tests were also made with the projected image of a continuously run moving picture film whose frame speed was synchronized with the vertical frame frequency of the reproducing equipment. This second test closely simulates the actual working condition of the tube if allowance is made for the higher line and frame frequency.

The first tube tested was found to be very insensitive, so much so that it was very difficult to distinguish the light image from the spurious image due to the line structure. There was evidence that even if a stronger signal could be obtained, the resolution would not be adequate. The cause of this loss in resolution appeared to be due partly to leakage and partly to redistribution losses. No estimate could be made of the definition on the barrier-grid side.

The second tube was prepared in such a way as to give a higher sensitivity and somewhat lower leakage. On test it was found that at high light levels, where the signal was well above noise and other spurious effects, the definition was very poor. As the light level was lowered the definition improved and at extremely low values of signal the definition approached that established by the line structure. Again leakage appeared to be the major cause of loss of definition with redistribution contributing an increasing amount as the light level was increased. The sensitivity of the barrier-grid side was very much lower than when the barrier grid was absent. Electric leakage between the elements and the barrier lines was responsible for the low response. In fact, this leakage was so high that the grid could not be made sufficiently negative to exclude all redistribution.

23.3 CONCLUSIONS

The tests outlined indicate that line-mosaic pickup tubes as constructed are not satisfactory from the standpoint of definition for the specific reconnaissance television application. The primary reason for this is leakage between elements where there is no barrier grid, and between the elements and the barrier grid with the barrier-grid mosaic. Redistribution also decreased the definition when no barrier grid was used.

This leakage probably is not fundamental. However, to overcome it a great deal of research on methods of making the line mosaic would have to be carried out.

A similar amount of research on a tube using a two-dimensional mosaic would also lead to a tube which could be used for the required reconnaissance. Such a tube would inherently be much more sensitive than any line-mosaic device, and the possible greater size of the former would be compensated by the simpler optical system required. Therefore, if a more extensive research program is to be undertaken to develop a reconnaissance television system, it would be better to concentrate on the working out of a two-dimensional mosaic pickup tube.

Chapter 24

SOUND RECORDING ON MAGNETIC MATERIALS

Development of a small spring-driven magnetic recorder-reproducer, in size about that of a 16-mm magazine-type motion picture camera, weighing 6 lb, giving telephone quality, and capable of recording in any position for a total time of 30 minutes. Further research should include reference to the final report on Project 13.3-87,[2] which gives a general survey of recording equipment available at the time.

24.1 INTRODUCTION

At the initiation of Project C-69[a] there was no lightweight portable pocket-sized recorder, in spite of the fact that all the different methods of recording — optical, mechanical, and magnetic — had been well developed.

The completion of the project[1] was marked by the production of a model which weighed 6 lb, including amplifier, batteries, microphone, and earphone. The recording medium was a stainless-steel wire 0.006 in. in diameter from which the recording could be erased when desired whereupon another recording could be made. The instrument can be attached to an auxiliary drive unit which operates from any 12-volt supply, such as the battery of a jeep or tank. This auxiliary drive eliminated the burden of winding the spring and gave more audio power on reproduction so that a number of observers could listen to the message simultaneously.

The work was originally undertaken to provide reconnaissance men with a means of recording what they saw and as an aid to memory and to preserve the information in case the scout was a casualty.

The development was used by government services as a basis for specialized designs. Much of the experience gained was successfully incorporated in magnetic recording equipment of different construction.

24.2 ADVANTAGES OF WIRE RECORDING

In preference to other means of recording, magnetic wire recording was chosen for the following reasons:

1. Mechanical vibrations do not seriously interfere with the recording or reproducing processes.
2. In any recording process in which space is a factor, the amount of recording medium used must be kept to a minimum. The space occupied by wire is a minimum for any required time of recording, as compared to that required by other media.
3. Little power is needed to make a recording.
4. The signal can be reproduced many times without deleterious effects on the recording.
5. The signal can be erased and the medium reused.

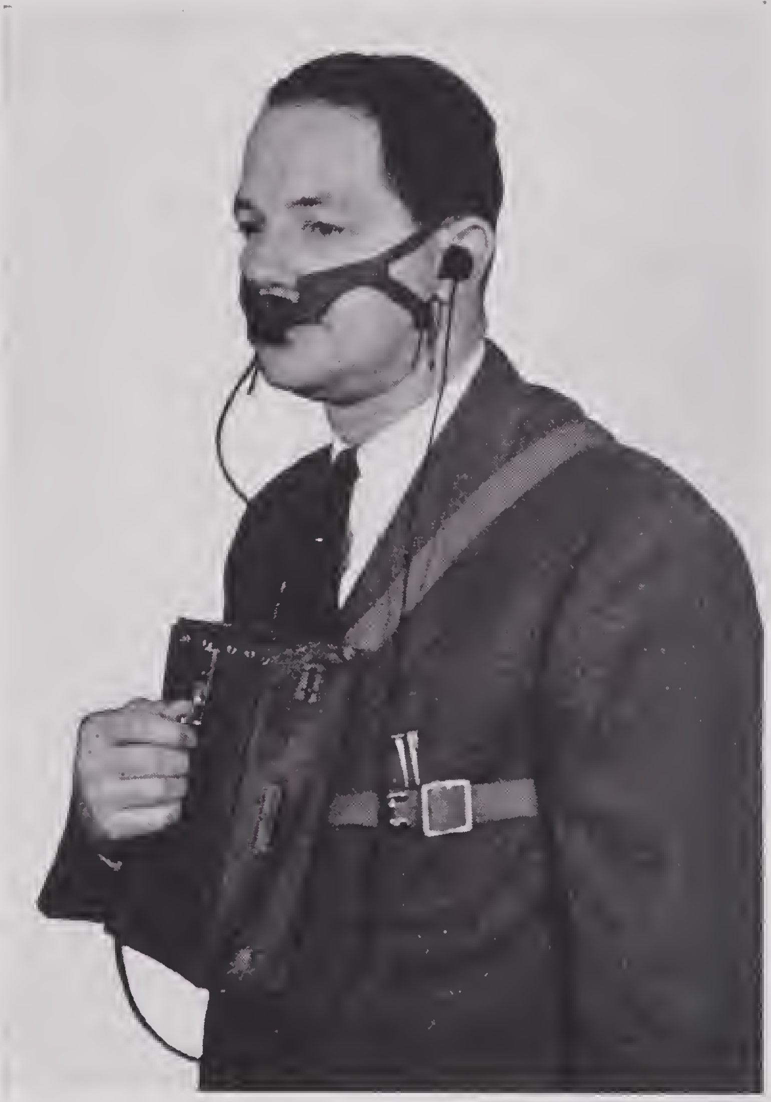

Figure 1. Portable recorder in use.

24.3 THE RECORDER

To eliminate the weight of batteries to supply running power for the reels, a spring-driven mechanism of the motion-picture camera type was selected.

A test setup was built consisting of two reels (take-up reel and supply reel), an electric-motor drive, and a magnetic ring head attached to a level-winding mechanism. This was used to determine what mechanical features were to be desired. From these investigations it became clear that the best method of keeping the wire under continuous tension was to control the starting and stopping by means of a brake applied to the supply reel. It also was found to be desirable to provide a winding-shaft spring in parallel with the mainspring which would be wound by the mainspring when the drive mechanism is triggered to stop.

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[a] Project C-69, Contract OEMsr-833, The Brush Development Company.

The level-wind device (similar to that used in a fish-line reel) was employed to obtain reasonably level winding of the wire. Level winding is necessary to prevent tangling of the wire on the reel and also to give good feed through the head with a minimum of vibration. This is advantageous in as far as background noise is thereby reduced.

Several 16-mm cameras were secured and studied to derive the drive mechanism. A spring was finally selected from an Eastman Kodak Company Model K camera and was incorporated in the experimental model, designated as D-103.

The mechanism contains a crank-wound spring which drives a gear train to the take-up reel. The reel rotates at about 350 rpm. This spring cannot be wound while driving the unit since it is of the type which has one end of the spring held stationary. The free end of the spring rotates in one direction when being wound and in the opposite direction when driving the gear train. This method of winding permits automatic stopping of the drive mechanism before the spring is completely exhausted and assures relatively constant speed of the take-up reel during its operating cycle. An alternative spring system was proposed which would enable continuous winding, but could not be worked out in a small space to include the advantageous feature of automatic shutoff when the spring is nearly exhausted.

Associated with the reel gear train is the governor gear train which operates the governor at a higher speed. The power is applied to the take-up reel through a friction clutch. This maintains synchronism of the level wind in both the rewind operation (when the supply reel is driven) and also for recording or playback. In this arrangement this level wind shaft winds wire on the take-up reel with a constant pitch of about 140 to the inch. Due to the variable speed of the supply reel with reference to the take-up reel the pitch with which the wire is rewound on the supply reel varies between 90 and 140 to the inch. Such variations in pitch have not been found to interfere with the proper operation of the equipment.

The recorder is controlled by a single lever which releases the brake acting on the supply reel and the winding shaft spring, putting the unit into operation. Operating time, starting with the main spring fully wound is approximately 30 seconds. The reel contains sufficient wire to record for a total period of 30 minutes. The reel diameter is 3 in. and its width ¾ in.

A feature of the portable recorder is the incorporation of three automatic stops. There is a worm-driven lead screw and nut which is timed so that when all the wire is wound on the take-up reel it automatically trips the control lever into the off position. When the portable recorder is used in conjunction with an external drive, the same nut operates a miniature switch to open the power supply for the external drive thus insuring that the wire will not be pulled off the reels on either playback record or rewind.

Provision was made to use the crank for winding the spring and rewinding the wire. The reel shafts and the spring shaft were fitted with splines to facilitate operation by the hand crank as well as by an external drive.

24.3.1 Wire Details

Wire 0.006 in. in diameter was decided on as a compromise between mechanical strength and a reasonably small storage factor. Magnetic qualities were determined by running a loop of test wire continuously over recording, reproducing, and erasing heads. It was determined that a wire speed of approximately 3 ft per second gave a frequency response of satisfactory intelligibility. Several wire samples were tried out and it was found that heat-treatment of these wires is needed for satisfactory signal-to-noise ratio. Carbon-steel wire properly heat-treated has a signal-to-noise ratio of 28 db or better. Stainless-steel wire later became available, with the advantage of corrosion resistance.

24.3.2 The Erasing and Recording Process

In this particular recorder the method of d-c erasing and d-c biasing is employed. This means that in the erasing process the wire is magnetically saturated. In the recording process a d-c field considerably smaller than the erasing field is superimposed upon the signal current. The biasing field is of opposite direction to the erasing field and chosen so as to select a zone in the hysteresis loop of the wire which for the signal amplitude is essentially linear.

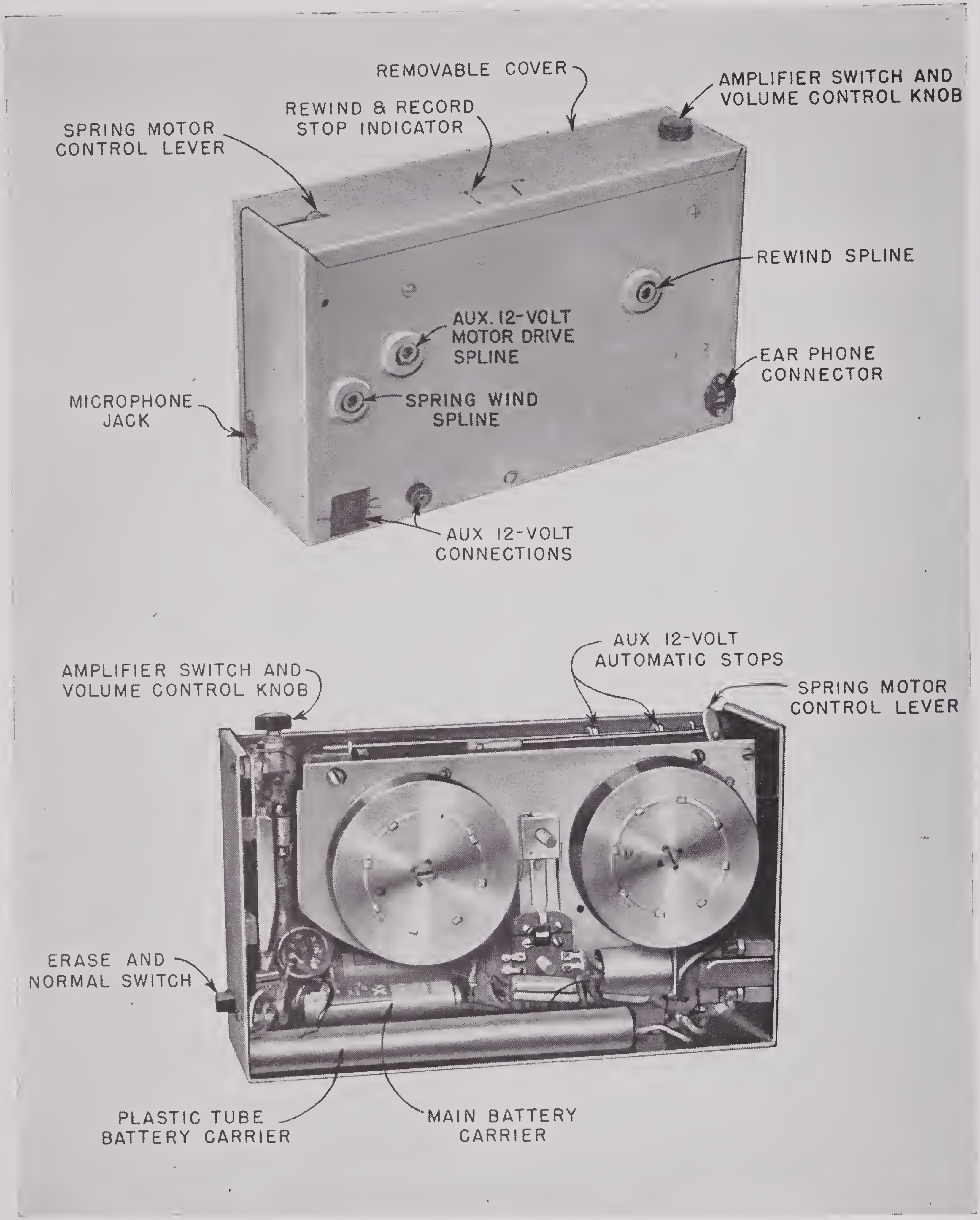
REMOVABLE COVER
AMPLIFIER SWITCH AND VOLUME CONTROL KNOB
REWIND & RECORD STOP INDICATOR
SPRING MOTOR CONTROL LEVER
REWIND SPLINE
AUX. 12-VOLT MOTOR DRIVE SPLINE
EAR PHONE CONNECTOR
MICROPHONE JACK
SPRING WIND SPLINE
AUX 12-VOLT CONNECTIONS
AUX 12-VOLT AUTOMATIC STOPS
AMPLIFIER SWITCH AND VOLUME CONTROL KNOB
SPRING MOTOR CONTROL LEVER
ERASE AND NORMAL SWITCH
PLASTIC TUBE BATTERY CARRIER
MAIN BATTERY CARRIER

Figure 2. Construction of recorder.

24.3.3 Recording Head

The recording head is constructed of two L-shaped Mu-metal pole pieces. The air gap over which the wire passes produces the leakage field necessary to generate the longitudinal magnetic recording pattern. A shoe is provided to guide the wire as it passes through the head, thus securing good magnetic contact.

The recording head is wound with 1,000 turns of No. 40 wire having a d-c resistance of about 70 ohms. The head has an inductance of 16 mh and a *Q* of 1.2 measured at 1,000 cycles per second. The erase current is 20 ma, and the biasing current is 7 ma. The maximum signal current can be 1 ma, with resultant undistorted magnetic impression. The unequalized power requirement for a signal of 1,000 cycles is of the order of ½ mw. To obtain better h-f response, equalization must be provided in the recording channel, which increases the power requirements.

The recording circuit consists of a carbon microphone (Type T-45), a 4½-volt "recording" battery to supply the current for the microphone, a transformer, and a resistor. The battery also supplies current for polarizing and for erasing the wire. A switch is provided which in its "normal" position (not "erase"), prepares the circuit for recording or playback. The microphone plug has to be inserted into its jack to close a miniature switch which controls the signal and polarizing currents.

The three cells, used for pen-sized flashlights, supplying the 4½ volts for the recording circuit were good for two hours of use.

The playback level obtained from the head is about 1 mv at 1,000 cycles per second.

24.3.4 Self-Contained Playback Amplifier

Because of the low output level, some means of amplification is necessary during reproduction. A hearing-aid amplifier was found to be suitable and is employed for this purpose. The three-stage amplifier is connected to the head by a small transformer designed to match 200 ohms to 1 megohm. Special care had to be taken to prevent microphonics, since these amplifiers are not normally intended for use in connection with rotating equipment. Rubber isolation proved successful.

24.3.5 12-Volt Auxiliary Drive

A separate semiportable motor-driven playback-rewind-erase unit was designed to operate as follows: The portable recorder can be clamped into position, where it automatically makes the proper mechanical and electrical connections. The auxiliary drive consists of a 12-volt d-c motor, three-tube amplifier using a vibrator as B supply, crystal microphone, earphones, and controls. It is contained in a plywood cabinet 10x16x17 in. and weighs about 30 lb. Power consumption is approximately 4 amp at 12 volts.

In using the two units together, the spring mechanism is by-passed. This assembly of two units can be used for all functional operations.

PART VIII

MISCELLANEOUS STUDIES

Chapter 25

SUBSTITUTES FOR NATURAL QUARTZ FOR FREQUENCY CONTROL

25.1 STATE OF THE ART

Prior to this project, the art of high-frequency control by means of quartz-crystal sections had been highly developed, but no synthetic (man-made) crystals for frequency control purposes had been found with the exception of Rochelle salt, which is not suitable on account of its chemical instability and extreme frequency variations with temperature. Since the sources for high-grade quartz crystal suitable for oscillator use are almost entirely outside the United States, it appeared most desirable that a way be found to manufacture frequency controlling elements from domestically available raw materials.

Project C-29[a] had the purpose of surveying possible lines of attack for this problem.

25.2 WORK ACCOMPLISHED

By a survey of available information, classification of such information, and some original theoretical work, the conclusions were reached that the only suitable substitute for natural quartz for frequency control in the conventional r-f range would be synthetic piezoelectric crystals and that there was a chance of finding such crystals, both among materials requiring high-temperature methods for crystal growth, and water-soluble materials. A number of crystal substances were proposed whose properties appeared promising. Some preliminary measurements were made.

The final report[1] was of a survey nature and no data on temperature dependence of resonant frequency, which is the most critical characteristic for the purpose considered, could be obtained within the scope and the time limit for the project.

Easing of the quartz supply situation made continuation of work on synthetic crystals for frequency control unnecessary at the particular time. As a long-range project, however, the Signal Corps has continued its interest and the contractor plans to carry on the work begun under Project C-29.

The final report contains an extensive bibliography (180 items) on frequency instability in oscillators, methods of increasing stability, quartz resonators, etc. In addition, the tables in the final report give much data on substances considered as possible quartz substitutes.

25.3 CONCLUSIONS

The only really promising way to replace natural quartz for frequency control in the range of 2 to 50 mc is to provide a suitable artificial piezoelectric material. Only thickness-controlled plate resonators vibrating in a shear mode and preferably suspended in corner- or edge-clamped holders come into question. In order to avoid length- or width-controlled parasitic vibrations, one of the outstanding difficulties at elevated frequencies, only crystals belonging to certain classes and only a few specified cuts are recommendable. The simultaneous achievement of freedom from parasitics and of highly reduced temperature dependence is still problematical. Substances in which crystal lattices are spatial frameworks of strong chemical bonds are required. Associated general properties, mostly necessary, partly highly desirable, are high melting point, chemical stability, good mechanical properties, low mechanical and electrical losses, and high mechanical wave velocities; good growing habits, i.e., possibility of obtaining faultless single crystals of about centimeter size, and fair piezoelectric coupling coefficient of the selected cut are also necessary.

Only high-melting crystals can be fully equivalent to quartz for application to high-frequency control. From the known crystal-chemical properties of beryllium oxide, it is concluded that this substance might yield useful high-frequency shear plates, possibly even superior to quartz plates. Several leads for an attack on the problem of obtaining beryllium oxide in large single crystals are available. Artificial growing of large quartz crystals is a problem of a similar nature. The advantage in case of quartz would be that all physical properties and parameters are well known and that it would represent the most direct replacement of the natural product. Aluminum orthophosphate and aluminum metaphosphate, which are related to quartz in their crystal structures, may have similar physical properties but seem to be easier to grow.

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[a] Project C-29, Contract No. OEMsr-120; Brush Development Co. This summary was written from the final report[1] on the project and from a letter from A. L. Williams, September 12, 1945, citing opinions of Jaffe and Baerwald, who did the work under the project.

Certain water-soluble crystals offer some promise as substitutes for quartz, with some sacrifice in the precision, constancy, and highest frequency attainable. The most hopeful crystals are $KLiSO_4$, $NaClO_3$, $NaLiSO_4$, $BeSO_4 \cdot 4H_2O$.

Production of large crystals of high-melting substances appears as an arduous and time-consuming but hopeful task of pioneering nature. Development work and establishment of a manufacturing process for the water-soluble crystals could proceed along established lines of research and engineering.

Chapter 26

SHIELDING FOR DIATHERMY

A study of methods of reducing radio interference from r-f power equipment used for noncommunication purposes.[a] Tests proved that if adequate filtering were provided to prevent coupling of the r-f energy to the 60-cycle power lines a single shielded room of 16-mesh bronze screening will provide an attenuation of approximately 73 db and that adding another shield (thus making a double-shielded room) will add another 73 db. In the final report[1] several types of line filters and several methods of constructing shielded rooms are described.

26.1 THE PROBLEM

The magnitude of the interference caused by diathermy or other noncommunication r-f power equipment at any receiver depends on several factors, notably the band width covered by the radiation and the power radiated, the geographical separation and the frequency separation between the source and the receiver being jammed, and the susceptibility of the receiver.

The band width of most diathermy equipment is quite large, because raw alternating current is applied to a self-rectifying oscillator. The use of filtered direct current on the plate of the oscillator appreciably reduces the influence of the radiating source by reducing the band width of the power radiated. However, such a scheme adds to the complexity and cost of the diathermy apparatus.

The load circuits into which diathermy apparatus works are of two general types, one being made up of two large insulated flat metallic plates separated by the body of the patient under treatment and the other consisting of a coil of a few turns of insulated flexible conductor placed near the portion of the patient's anatomy which is to be treated. In either case the load system represents a transmission line system with leads rather widely separated and with rather high voltages across them. Furthermore, the leads acting as a transmission line may be an appreciable fraction of a wavelength long and, not being terminated in their characteristic impedance, may have standing waves upon them. Thus a considerable radiation of energy may take place, since the load and the leads connecting the machine to it constitute a fairly good antenna system.

Eliminating the radiation from the leads or the load circuit is not of much use unless the radiation from the oscillator itself is eliminated; this implies adequate shielding of the whole system, which can best be accomplished by working the whole system in a shielded room.

26.2 TESTS CONDUCTED

In the study of means of preventing troublesome interference the several components of a typical oscillator-amplifier equipment were separated and individually shielded.

It was found that the use of balanced circuits such as push-pull oscillators or balanced load lines will reduce the interference materially.

The effect of grounding a shielded room was also investigated. It was ascertained that if all joints in the shield, such as doors and windows, were tight and if the filter used in the power line were adequate it made no difference in the attenuation obtained whether the system were connected to an external ground or not.

It was found that an *LC* filter in which special pains are taken to keep the condenser leads short constitutes an excellent filter, provided it is enclosed in an electrically tight box and bonded directly to the shielding where power enters the room. Special pains must be taken to build and install a filter which will allow the full attenuation available to be obtained from a double-shielded room. Shielded transformers of conventional design were found to be inferior to *LC* filters.

The final report gives the results of the individual efforts made and describes the shielded rooms and filters constructed and gives cost figures. The cost of reducing interference by shielding was approximately $3.00 per db.

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[a] Project C-31, Contract OEMsr-225, Rensselaer Polytechnic Institute.

Chapter 27

LOCATING FAULTS IN WIRE LINES

By a special bridge located at a terminal point, capacitance changes between conductors provide means for determining the location of an open fault in W110B twisted pair and WC548T1 "spiral four" lines. A keyed oscillator located at the terminals and a portable amplifier carried along the wires provide additional means of locating the open point.

27.1 THE PROBLEM

The object of Project C-37[a] was to find a method of determining the point of occurrence of open faults in wire lines usually of the W110B twisted pair, and WC548T1 four-wire spiral armored-cable types. It was desired that the fault be located first by measurements from a switchboard and then that some method be provided so that a lineman might proceed along the line with an instrument which would give an indication of the exact location of the break.

Several suggested methods of attack were explored but all basically were methods of measuring the capacitance between wires or from the open wire to ground. The best solution, therefore, appeared to be the accurate measurement of the capacitance of the wire to some nearby object.

27.2 CAPACITANCE MEASUREMENTS

Crude methods of open-fault locations by means of voltmeter readings and a-c readings have existed for many years, but their accuracy depends upon weather and line conditions. Bridge methods are much more accurate, yet the accuracy with which the length of a wire to an open fault may be measured depends greatly on which of several capacitances is measured. Where a wire or pair of wires is suspended above the earth but the position with respect to the earth is alterable, either by a change in distance or by a change in the apparent ground level due to rain or other cause, the capacitance value also varies. For this reason, the least variable of the capacitances associated with a wire is that between the wire and its mate in a pair or the other conductors of a cable.

To measure this capacitance, however, it is necessary to remove the effects of other capacitances which would otherwise vitiate the results. This can be done.

Referring to Figure 1A, a pair of wires, 1 and 2, are shown together with their capacitances to each other and to the ground. If C_{1g} is the capacitance of 1 to ground, C_{2g} that of 2, and C_w the capacitance between wires, and if the wires are connected to a capacitance bridge as shown, with wire 2 grounded, the measurement of capacitance (known as C_1), will be $C = C_{1g} + C_w$; the grounding of wire 2 has removed C_{2g} from the measurement. If a second measurement (C_2), as in Figure 1B, is made with wire 1 grounded,

$$C_2 = C_{2g} + C_w.$$

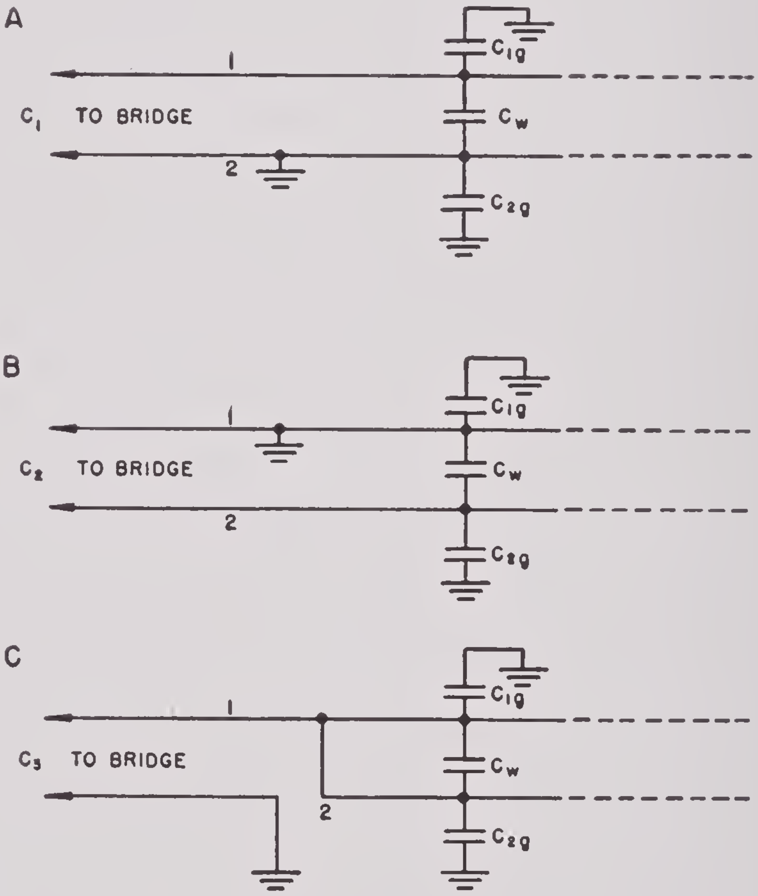

Figure 1. Several capacitances existing in pair of wires.

A third measurement C_3, is now taken, as in Figure 1C, with wires 1 and 2 connected together, and

$$C_3 = C_{1g} + C_{2g}.$$

We now have three simultaneous equations in three unknowns, so that by combining

$$C_w = \frac{C_1 + C_2 - C_3}{2},$$

C_w is the actual capacitance between the two wires.

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[a] Project C-37, Contract OEMsr-316, Western Union Telegraph Co. Detailed drawings contained in the original project report[1] showing mechanical layout of these instruments have been omitted from this summary.

This value should be recorded in the switchboard records as soon as the lines are installed. This value should be checked from time to time.

27.2.1 Errors Due to Resistance and Leakage

An error is introduced in the measurement by the effect of the resistance of the wires in series with this capacitance but this is eliminated by a modification of the bridge whereby S in the bridge arm (Figure 2) balances out the series resistance of the wires. The line makes up one arm of the bridge and, since the two resistive arms are equal, the balancing arm S and C is direct reading. Leakage between wires or between either wire and ground would also cause a measurement error and for this reason the bridge has been further modified as in Figure 3. The method used is first to apply direct current to the bridge, adjusting rheostat P for a null on the galvanometer G. This is a d-c balance and the result is that the sum of the resistance "in circuit" in P plus the total resistance of the potentiometer is equal to the sum of the series line resistance R_s in line 1 plus the leakage resistance R_p of line 1.

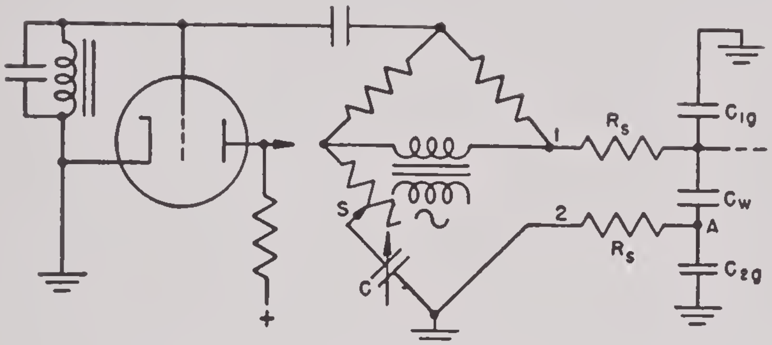

FIGURE 2. Capacitance bridge in which S balances out series resistance of line.

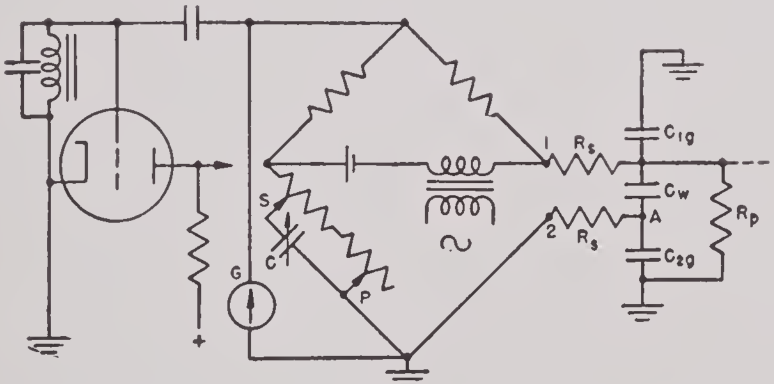

FIGURE 3. Further modification of elementary capacitance bridge to compensate both wire resistance and leakage.

The position of the arm of potentiometer S will not affect the d-c balance. Rheostat adjustment is left *untouched* from this point on during the measurement of C_1; S and C are alternately adjusted for the most complete null. C_2 and C_3 are measured by the same procedure, the a-c balance in each case being preceded by the d-c balance. The final adjustment of the bridge will be such that the resistance across C will equal the line leakage resistance R_p; the remainder of the resistance in S will equal the line resistance R_s. The capacitance setting of C is still the line capacitance.

This procedure makes possible the measurement of capacitance with any combination of line leakage and line resistance within reason, for all of the line connections shown in Figure 1.

27.3 PORTABLE LOCATOR

The second part of this project was to find means by which a lineman might proceed parallel to the line containing the open fault and obtain an indication of the location of the fault without placing too much dependence upon the switchboard measurements and calculations made therefrom. This would be particularly applicable in the case of W110B twisted pair, where the capacitance change is as great as 5 per cent.

By energizing the line under test with alternating current at around 1,000 cycles, an alternating electrostatic field is set up around the wire or pair. This may be detected on using a small antenna and a multistage amplifier, battery operated. Under favorable conditions, depending upon the shielding effect of other wires, the voltage of the test frequency, the height of conductor above ground, and noise from sources of interference and on open wire, the signal has been detected at points 2,000 ft from the line using a 50-volt signal to line and an amplifier with a sensitivity such that a 160-μv input gives full output.

If there were only one wire in the line such an arrangement would suffice and it would be necessary only to follow the line in a car on which an antenna had been attached, going out to the point where the signal disappeared. With other wires in close proximity, however, sufficient energy would probably leak past the break to give a poor indication of the fault.

To provide better discrimination, the arrangement shown in Figure 4A may be used. The faulty wire and a good wire (preferably the second wire of the pair) are connected to the secondary terminals of a transformer which has its mid-point (secondary) grounded. The two wires are thus carrying voltages 180 degrees out of phase with each other and, if the electrostatic

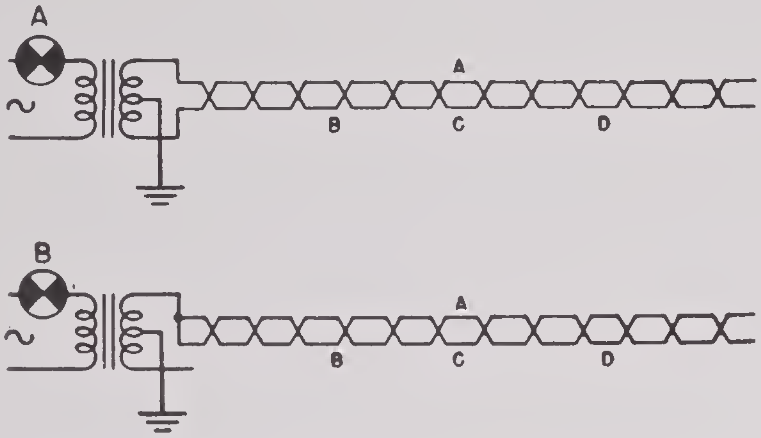

Figure 4. Switching cycles to make easier location of an open fault at position A.

field at point *B* is investigated with the portable receiver, there will be very little signal, since the two will cancel each other almost completely. As we go through point *C* to point *D*, the signal increases, due to the fact that only one wire is involved at this point, therefore no cancellation results. The transition from point *B* to point *D* is very distinct. The change can first be noticed when the distance from the antenna to the break is about twice that from the antenna to the wire. While the above method is an improvement on the use of a single wire and permits the open fault to be located in the presence of other wires, the degree of cancellation of the two out-of-phase voltages may not be complete enough to avoid confusion. If, however, the two wires are connected as in Figure 4B, for, say, ¼ second and then connected in phase opposition for a like period, we will have a condition where on the near side of the fault a signal will be received which is strong for one period and weak, if audible at all, for the second period. On the far side of the break, the signals will be approximately equal in strength.

This is shown in Figure 5. On the near side a signal similar to Figure 5A is received in which alternate signals are weak or inaudible, becoming equal after the open fault is passed. Thus, a lineman may, if necessary, only approach the line at intervals, yet he will immediately know from the character of the signal he receives whether he has passed the break. If it so happens that both wires of a pair are open, it is only necessary to use a third wire paralleling the two faulty ones as a good wire, the two faulty wires being strapped together to form one conductor. The switching operation and the source of alternating current have been combined in one instrument which is equipped with a power supply operating from 110 volts of alternating current. This instrument, known as an oscillator-keyer, contains a small motor-driven switch, a voltmeter indicates the power level, and two signal lights indicate whether power and motor are connected to the power source.

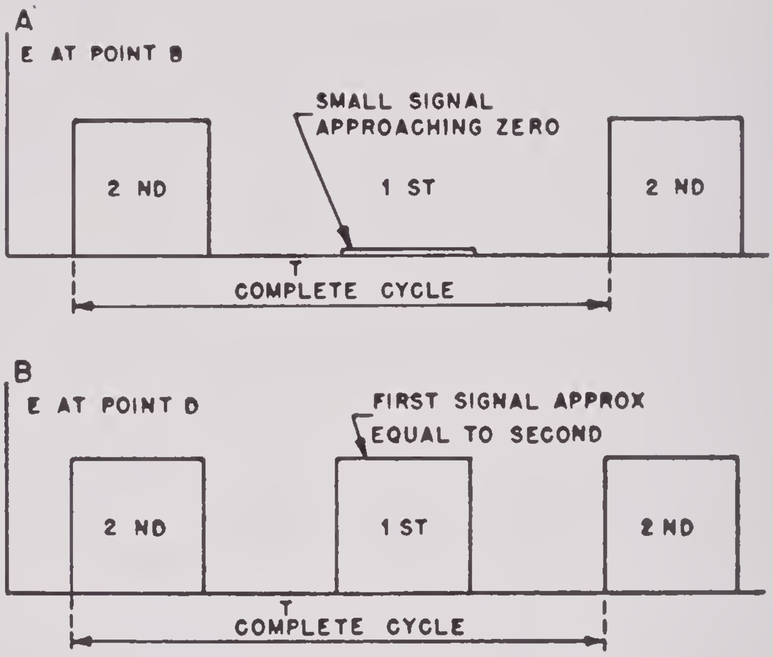

Figure 5. Effect of switching on electrostatic field near wire. In A, listener is at point *B;* in B, listener has passed fault and gets equal signals on both of switching cycles.

The portable locator is a multistage battery-operated amplifier. A small metal wand or an antenna, permanently attached to a car, may be plugged into a socket on one end of the instrument. The signals are noted by watching a rectifier-type meter in the output stage of the amplifier or by listening to the output of a miniature speaker.

In operation, the level of gain is raised to a point where full deflection of the meter occurs upon the reception of the signal. No harm results from excessive gain since the meter is protected by the limited power output of the driving tube.

Schematics of the capacitance bridge, the oscillator-keyer, and the portable fault locator are shown in Figures 6, 7, and 8.

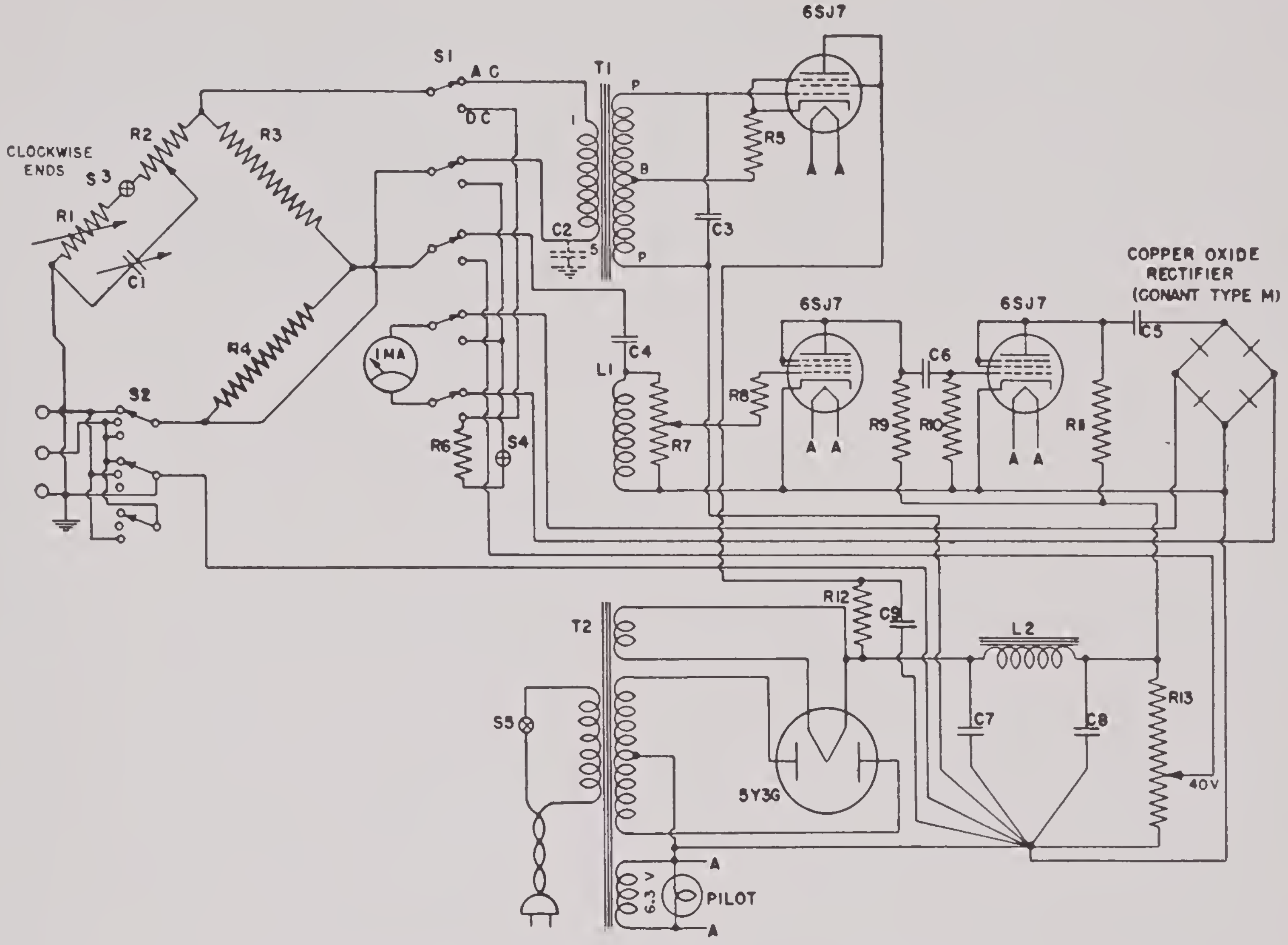

Figure 6. Schematic diagram of capacitance bridge.

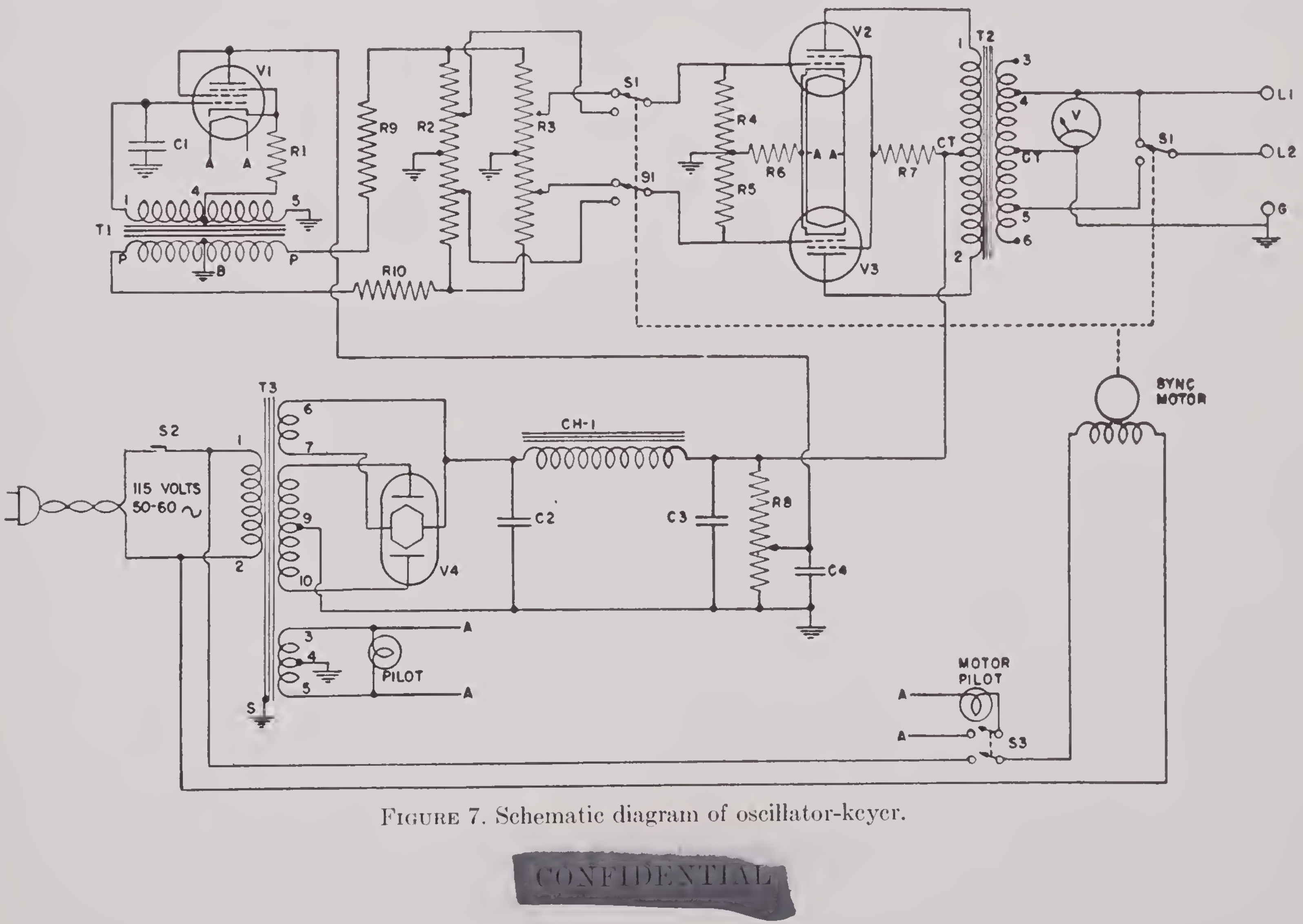

Figure 7. Schematic diagram of oscillator-keyer.

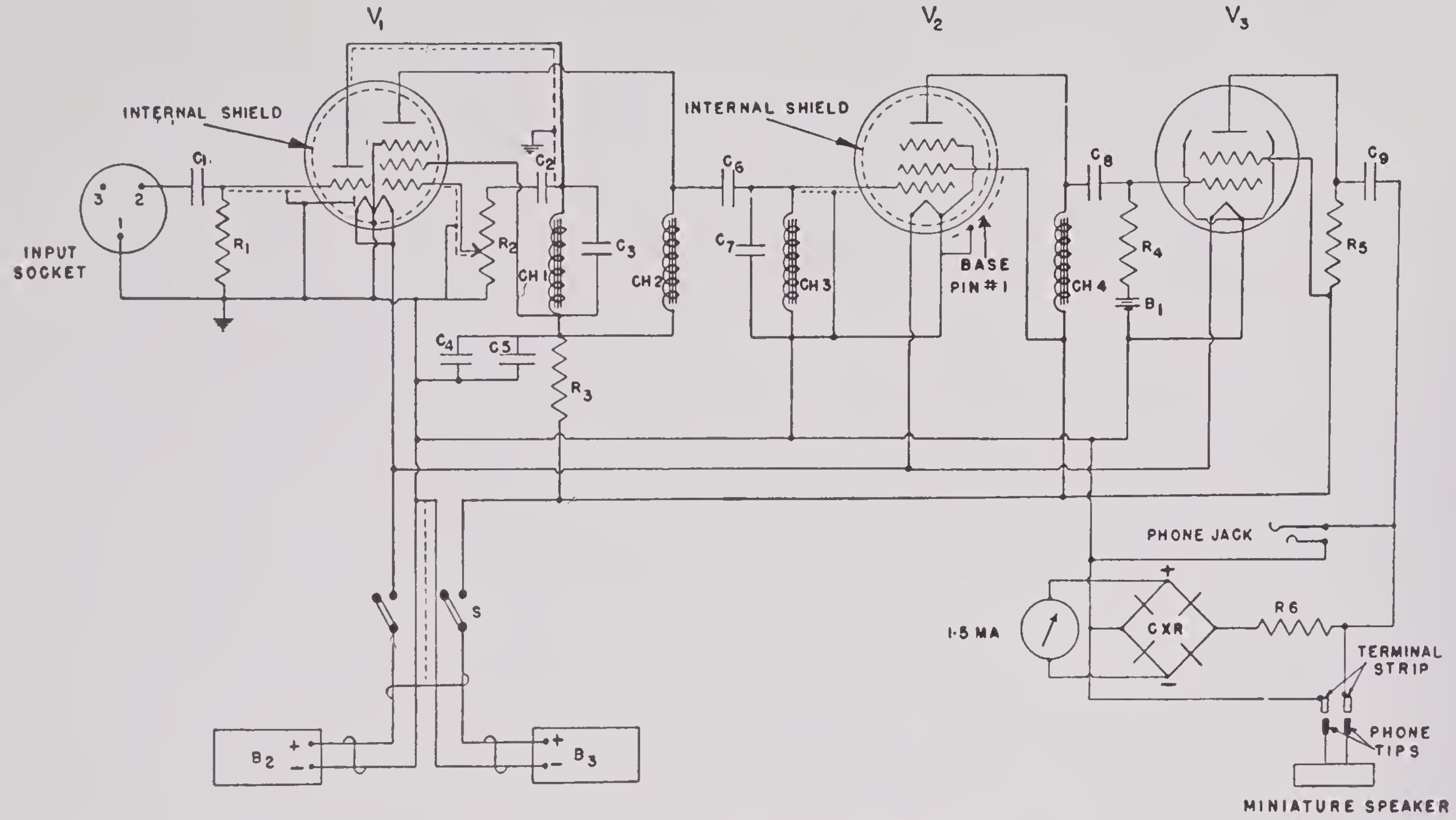

FIGURE 8. Schematic of portable fault locater.

Chapter 28

STORAGE BATTERIES FOR COLD CLIMATES

Research leading to the development of small storage batteries giving 40 to 46 per cent of normal, or 80-F, capacities at —40 F. Previous to this development, capacities of 19 to 25 per cent had been obtained.

28.1 STATE OF THE ART

The research covered by Project C-40[a] was considered for an appreciable period of time before the entry of the United States into active warfare. Studies of the conditions under which combatant forces were engaged, along with the rapid development of various types of highly specialized apparatus requiring self-contained power supplies, brought forth several important facts.

1. The performance of the primary cell (dry cell), although adequate for operations at normal temperatures, was entirely inadequate at low temperatures.
2. The performance of the lead-acid secondary cell, although operative at temperatures much lower than those at which dry cells failed to operate, was not generally satisfactory.
3. The secondary or storage-battery cell with adequate capacity to meet the operating conditions required at low temperatures and the fixed dimensions of the apparatus with which it would have to be used was not suitable because of its physical size.
4. Studies of low operating temperatures as recorded by the U. S. Weather Bureau at Fairbanks, Alaska, indicated that although extreme temperatures of —60 F might be met infrequently and for short periods of time, the temperatures during the winter months of November to March ranged generally between —30 and —40 F. Adequate capacity for operation at —40 F would not only be desirable but entirely necessary.

28.2 THE RESEARCH PROBLEM

The research problem resolved itself into two distinct operations or projects:

1. The investigation of the possibilities of producing small nonspill or leakproof storage batteries, similar in shape and dimension to the primary or dry cell, which would have satisfactory performance at a temperature of —40 F.
2. The refinement of design and the investigation of manufacturing methods leading to the production of satisfactory storage-battery cells which would meet the prevailing operation conditions.

28.2.1 General Plan of Attack

As the storage battery is a chemical device, the general plan of attack had to be one wherein the functions of each component part were considered and studied. The investigation of the individual component parts was carried out in this order:

Insulation. To obtain maximum capacity with a minimum of free electrolyte it was necessary to investigate new types of insulation. To accomplish this it was necessary to evaluate the electrolyte-retention values of various types of packed insulation as well as performance at normal 80 F and at —40 F. Materials such as glass wool, cellulose paper, Latex sponge, pressed wood pulp, and Fibrite were examined.

Grid Thickness. By the use of plates pasted with the same positive and negative mixes, the determination of plate combinations to obtain the best ratios between warm and cold capacities and the best plate thickness, surface, and contour was determined. Plate thicknesses of 0.04, 0.07 and 0.093 in. were tested. It was determined that the use of multiple thin plates was superior to the use of fewer thick plates.

Mixes. Having established the best experimental plate designs, various positive and negative mixes were investigated. This was carried out in cells having no insulation. Mixes superior to standard practice were developed.

Insulation. Tests carried out by the addition of various types of insulation to the cells built with the best mixes showed that five of the insulating materials used would not seriously affect the capacity at high or low temperatures.

Following these tests, work was done to determine the best grid design on the basis of conductivity and manufacturing facility, and on the physical and chemical properties of various container materials as well as their ease of manufacture and availability.

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[a] Project C-40, Contract OEMsr-420, Willard Storage Battery Co.

28.3 RESULTS ATTAINED

Final testing resulted in combinations where capacities of 40 to 46 per cent of the normal or 80-F capacities were obtained at —40 F. Previous to this development capacities of 19 to 25 per cent had been obtained. Considering the fact that primary cells are practically inoperative at —40 F or adjacent temperatures, and that the available capacity has been increased as indicated, the conclusion was reached that it was possible to develop the chemical and physical properties of small storage batteries to the point where satisfactory capabilities may be obtained at cold temperatures.

The design and development of production models was taken over by the Signal Corps Development Laboratories. Production models of the D and No. 6 sizes of primary cells and the addition of a new type DD or double-D cell were made under this program.

28.4 DEVELOPMENT OF A ONE-CYCLE CELL

After the initiation of the research described above, urgent demands were made by the several agencies for the exploration of single-cycle (one-shot) storage cells or batteries to replace the basic cell used in B batteries.

The exploration work accomplished under the early part of the project had led to the development of a satisfactory cell for use at —40 F. The element of this cell was made up of flat plates and took the conventional form of a rectangular prism.

To be interchangeable with the dry cell, the flat-plate element would have to be installed in a cylindrical container. Such installation would be expensive from the standpoint of the volume required, since a rectangular element in a cylinder would not use the volume efficiently. Capacity would be sacrificed.

Much work was done on elements of various shapes. For example, certain rolled elements, especially of the "herringbone design," indicated good possibilities, in fact better than any other investigated especially for installation in cylindrical containers. However, new and special operating machinery would be required to put this element into production. For this reason it was thought advisable not to recommend production of this type but to concentrate upon flat-plate cells in the rest of the work.

Data were collected and procedures developed which led to the production of a line of miniature sizes of lead-acid flat-plate type cells for nonspill, nonslop construction for either one-cycle or for repeated cycling service. The assembly of such cells into simple packs could be accomplished by merely assembling a pack of the individual cells or by assembling the elements into monobloc plastic containers, as was done in the Signal Corps 20-volt Type ER-B-20.

The utility of the single-cell unit lay in quick assembly into simple packs for nonstandard usage or where out-of-the-ordinary applications made it preferable to other types of cells.

Chapter 29

AN ELECTRICAL CANCELLATION AND INDICATING SYSTEM

A system which will cancel out any recurrent complex wave and give an alarm indication automatically when new wave components appear. Useful in monitoring systems, since, by establishing an alarm threshold, an alarm can be given when new signal or noise elements occur.

29.1 STATEMENT OF PROBLEM

The problem as presented in NDRC Project C-67[a] was to design an electrical system which would provide:

1. An adjustable complex voltage waveform which could be made very nearly identical to any repetitive wave that may be produced by an unknown source.
2. Automatic comparison of the known and unknown waves in such a manner that their resultant will be zero until a change occurs in the unknown wave shape.
3. An aural or visual alarm to indicate such change in the unknown wave regardless of the polarity of this change.

29.2 SOLUTION TO THE PROBLEM

After considering several possible methods described in the final report,[1] the following scheme was developed. Beginning with a repeating square wave of duration δ and period τ, of variable amplitude, any wave of the same fundamental period τ can be approximated by a series of pulses of equal duration and period, but so time shifted that the start of each coincides with the end of the preceding pulse (a time delay of δ seconds). (See Figure 1.)

Considering a composite wave made up of m such pulses with their durations adjusted to completely fill the period τ, i.e., $\delta = \tau/m$, the amplitude coefficient $A_{n,\alpha}$ for any one pulse α becomes

$$A_{n,\alpha} = \frac{2E_\alpha}{m}\left(\frac{\sin\frac{n\pi}{m}}{\frac{n\pi}{m}}\right)\cos\left[\frac{(2\alpha - 1)n\pi}{m}\right],$$

where $\alpha = 1, 2, 3, 4 \cdots (m - 1), m,$

m = total number of pulses in the repeating wave, and

n = order of the harmonic for a particular pulse.

These pulses have fixed durations and fixed phase displacements. The only variable for a particular pulse is its amplitude E. Moreover, each adjustment is independent and need be set only once. The cancellation procedure can thus be organized in any simple manner. For example, pulse No. 1 can be adjusted in amplitude to match the unknown wave for its portion of the total cycle, pulse No. 2 for its portion, etc. (See Figure 2.)

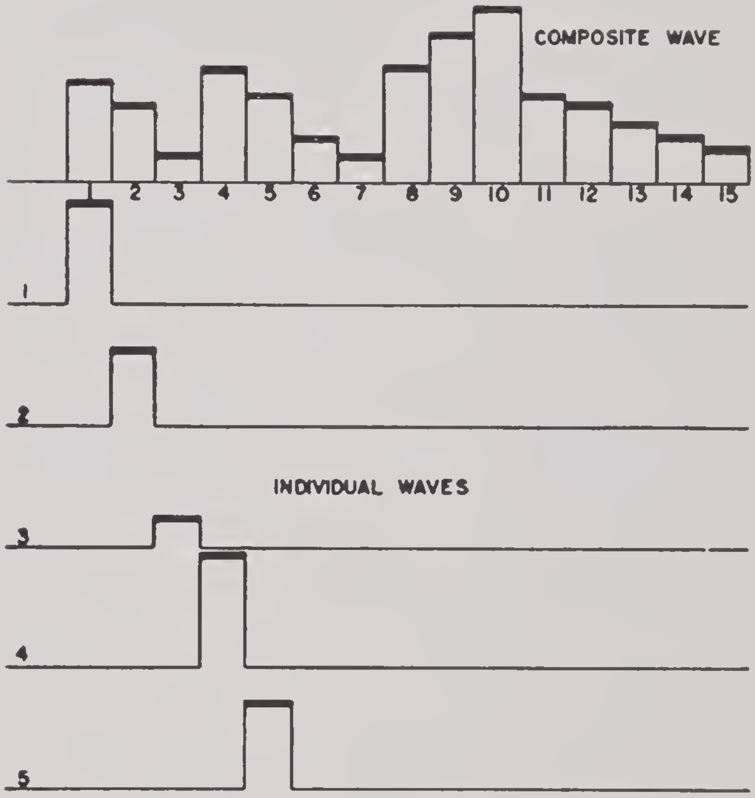

Figure 1. Composite wave made up of individual elements of varying heights, each occuring immediately after one preceding.

Figure 2 also illustrates the method of obtaining function 1. The polarity of the local or known wave is inverted and mixed with the unknown wave in a comparator circuit the output of which is proportional to the difference between the two waves. Actually, this output is a measure of the degree of approximation achieved and will improve with the number of pulses composing the synthesized wave. Using a Fourier schedule as a guide, it was decided to utilize 40 rectangular pulses in the experimental work ($m = 40$). With this value of m, Figure 2 illustrates only a portion of the wave period.

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[a] Project C-67, Contract OEMsr-748, RCA Laboratories. Only basic elements of the project are given in this summary. The analytical portions of the final report with necessary diagrams and circuits will be found in the report and on microfilm.

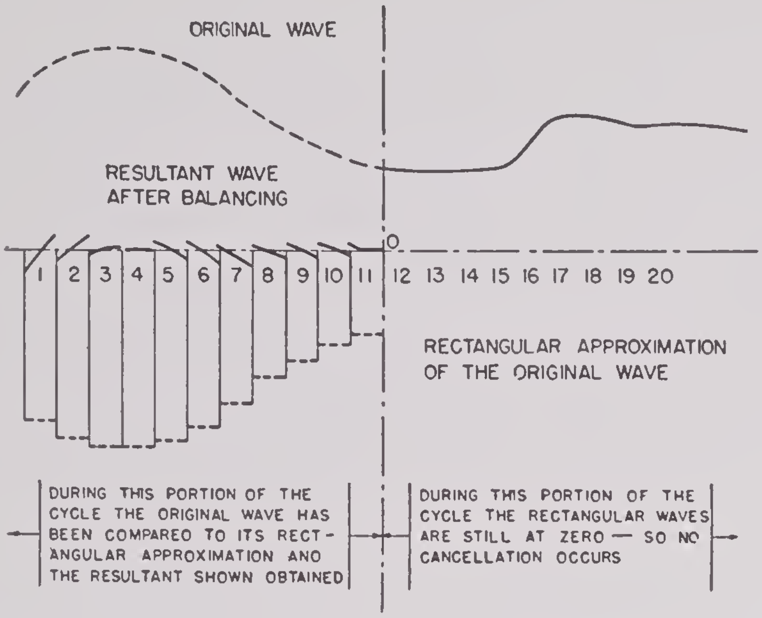
ORIGINAL WAVE
RESULTANT WAVE AFTER BALANCING
O
1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20
RECTANGULAR APPROXIMATION OF THE ORIGINAL WAVE
DURING THIS PORTION OF THE CYCLE THE ORIGINAL WAVE HAS BEEN COMPARED TO ITS RECTANGULAR APPROXIMATION AND THE RESULTANT SHOWN OBTAINED
DURING THIS PORTION OF THE CYCLE THE RECTANGULAR WAVES ARE STILL AT ZERO — SO NO CANCELLATION OCCURS

Figure 2. Partial cancellation of wave by opposing to it a series of rectangular waves.

In general, the rectangular pulse method of approximation will leave a triangular residual of noise as shown. This noise can be considerably reduced by shunting the synthesized electrical wave with an appropriate capacitance as illustrated in Figure 3. The degree of approximation is thus improved correspondingly.

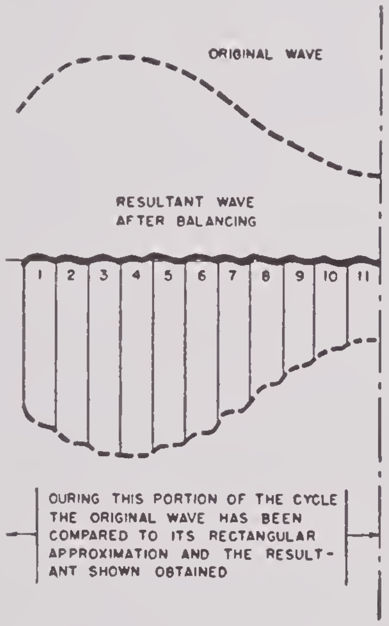
ORIGINAL WAVE
RESULTANT WAVE AFTER BALANCING
1 2 3 4 5 6 7 8 9 10 11
DURING THIS PORTION OF THE CYCLE THE ORIGINAL WAVE HAS BEEN COMPARED TO ITS RECTANGULAR APPROXIMATION AND THE RESULTANT SHOWN OBTAINED

Figure 3. Use of capacitance to further cancelling effect.

Having determined *m,* the number of pulses required in the local wave, consideration was next given to the method of their generation. One proposal was to use 40 trigger circuits so arranged that each would trip immediately following the preceding one. Such an arrangement is straightforward from a design standpoint, but the number of tube circuits involved is considerable. Consequently, this method was discarded in favor of a commutator and brush assembly requiring very little additional equipment.

Preliminary tests of the commutator brush-arm assembly as a complex-wave generator were carried out using a 42-segment face plate originally built for telegraph work. In the initial setup, each of 40 of these segments was connected to the movable arm of one of 40 potentiometers. The potentiometers in turn had their respective high-impedance terminals connected in parallel. The remaining two segments were used to provide external synchronizing pulses to a standard oscilloscope. A schematic diagram of the experimental test arrangement is shown in Figure 4. The angular velocity of the rotating brush arm in this case was approximately 15 cycles per second, thus requiring very good low-frequency response in the succeeding amplifiers. This angular speed could not be increased materially without causing noticeable splits between the rectangular pulses unless great care was taken in the adjustment of the brush position. It was determined experimentally that a better method was to tie diagonally opposite segments in parallel and provide twice the number of segments. The fundamental frequency would then be doubled for the same angular velocity since the local wave would be repeated twice for each rotation of the brush. In the same way, the frequency could be tripled with $3m$ segments, etc. The results of these tests indicated conclusively that the commutator method was the simplest and most satisfactory type of complex-wave generator.

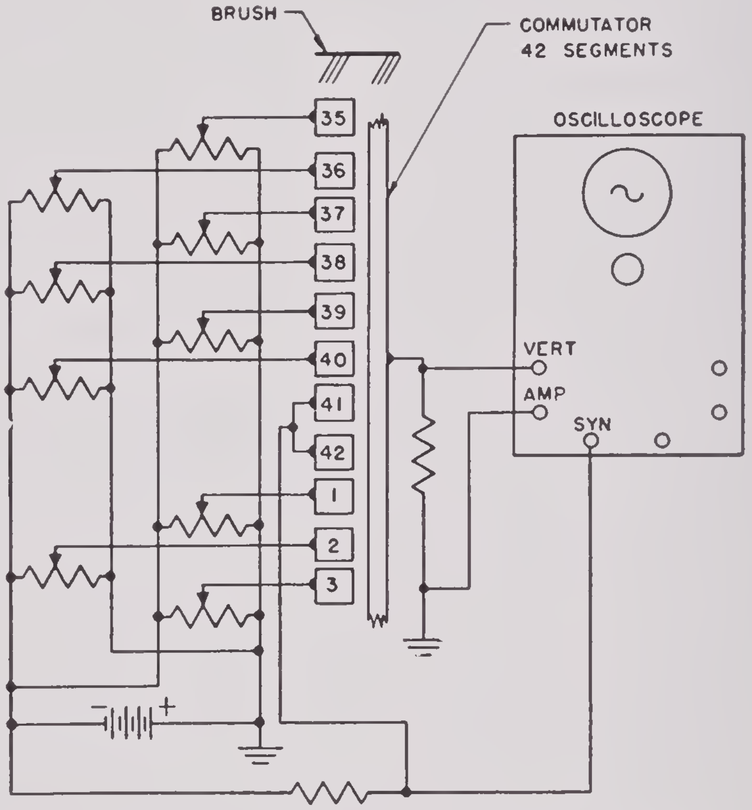

Figure 4. Elements of cancelling system employing rotating commutator to produce cancelling waves.

On the basis of these experimental tests, it was decided that the complete cancellation and indicating system should include:

1. A complex-wave generator of the commutator type.
2. An electronic mixer with linear input-output characteristics.
3. An oscilloscope to show the accuracy of approximation to the unknown wave.
4. An alarm system.

A block diagram of this system and its interconnections is shown in Figure 5.

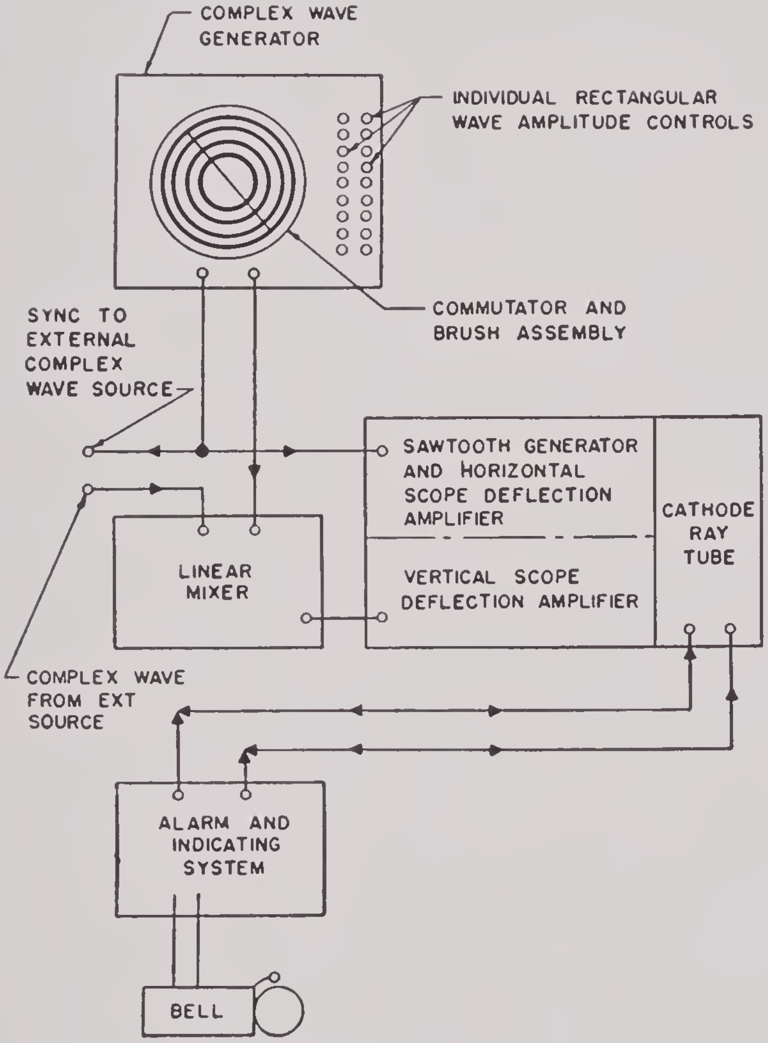
COMPLEX WAVE GENERATOR
INDIVIDUAL RECTANGULAR WAVE AMPLITUDE CONTROLS
COMMUTATOR AND BRUSH ASSEMBLY
SYNC TO EXTERNAL COMPLEX WAVE SOURCE
SAWTOOTH GENERATOR AND HORIZONTAL SCOPE DEFLECTION AMPLIFIER
CATHODE RAY TUBE
LINEAR MIXER
VERTICAL SCOPE DEFLECTION AMPLIFIER
COMPLEX WAVE FROM EXT SOURCE
ALARM AND INDICATING SYSTEM
BELL

Figure 5. Block diagram of system.

29.3 PROPOSED ELECTRICAL CANCELLATION AND INDICATING SYSTEM

To realize fully the advantages of the system shown in Figure 5, it was deemed necessary to design and experimentally verify each of the major circuits. Moreover, this would permit an accurate evaluation of the overall system from a practical standpoint. These circuits, as finally developed, are described below.

29.3.1 Complex-Wave Generator

The multisegment commutator was constructed in accordance with standard telegraph practice. Four concentric rings are provided in the arrangement, of which two are solid and two are segmented. The outside ring (No. 1) is divided into 84 equilength segments of which the first 80 are paired by interconnecting diametrically opposite segments. This was done to double the fundamental output frequency. Thus the complex-wave output, made up of 40 successive rectangular pulses, has a fundamental frequency of 30 cycles per second even though the brush actually rotates at 15 cycles per second. The angular speed of the brush is derived from an 1,800 rpm synchronous motor operating through a two-to-one step-down gear train. The remaining 4 segments (41, 42, 43, and 44) are connected to segment 40. During the time the brush is passing over these segments, a second brush on the same arm makes contact with the isolated segments shown on ring (No. 2) (see Figure 6). These segments form part of the sawtooth generator which, by periodically discharging capacitor C_{17} produces the sawtooth voltage used for oscilloscope deflection and external synchronizing purposes.

Mixer Stage. To utilize the cancellation device in various applications, it was necessary to insure linear operation throughout. The actual circuit is given in Figure 6. Since tubes V_1 and V_2 have nonlinear input-output characteristics, one must resort to graphical analysis for an exact solution.

Because of resistance elements common to both halves of V_1 (Figure 6), the mixer stage of the system, it was noted that the d-c operating conditions of the right-hand triode changed as the average grid potential applied to the left-hand triode varied. The question arose whether these d-c changes would adversely affect the mixer linearity over its operating range.

To determine this effect, it was necessary to obtain the magnitude of the shift in the operating point of the right-hand triode, as the average potential applied to the grid of the left-hand triode varied over its maximum range. Once this had been determined, assuming the input versus output voltage characteristics for the right-hand triode were drawn for the two extreme operating points, the effect of the shifting operating point would be known.

This problem was approached graphically, Figure 7 showing that linearity does exist in the mixer stage.[b]

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[b] The analysis of this circuit with the necessary voltage-current characteristics will be found in the final report.[1]

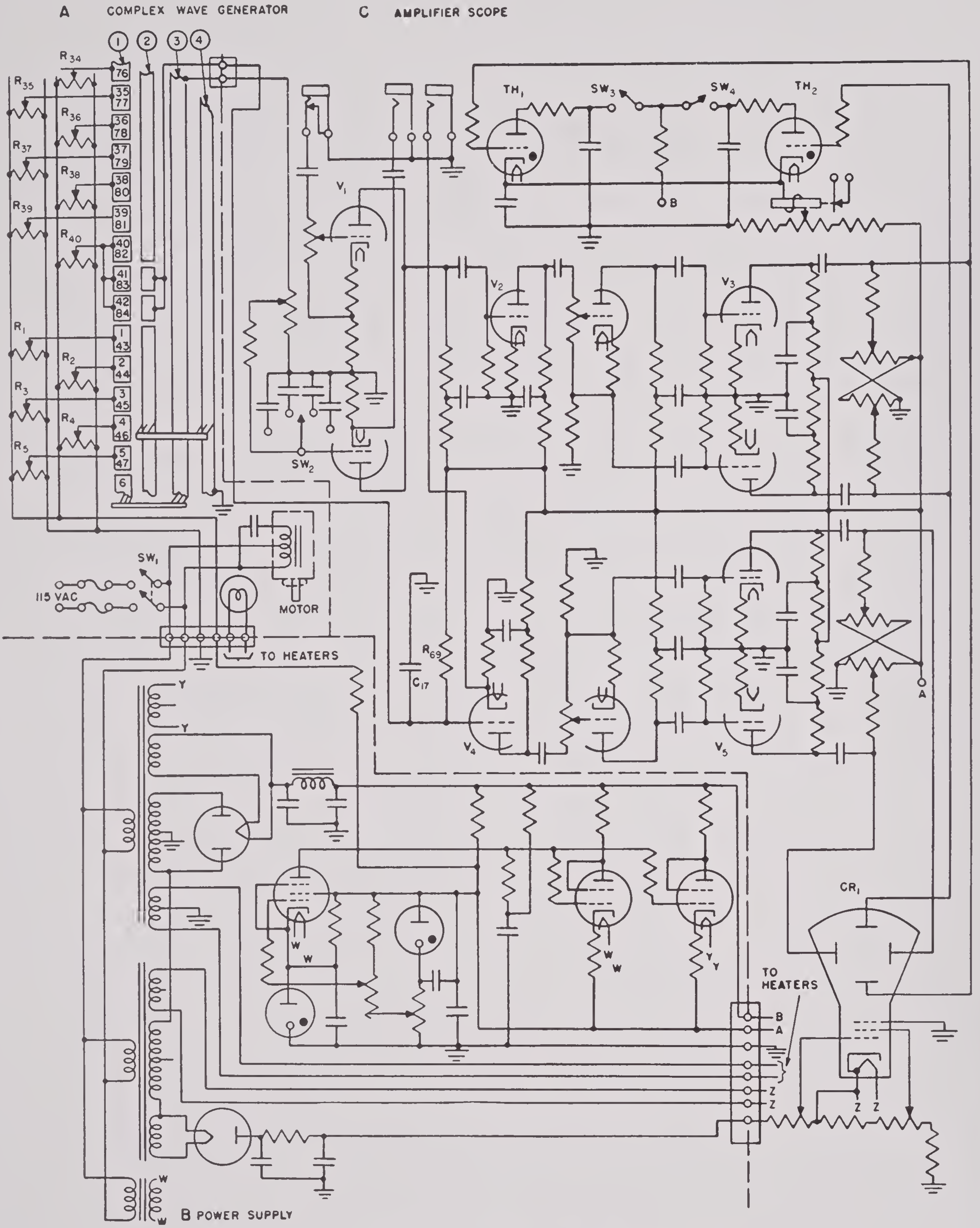
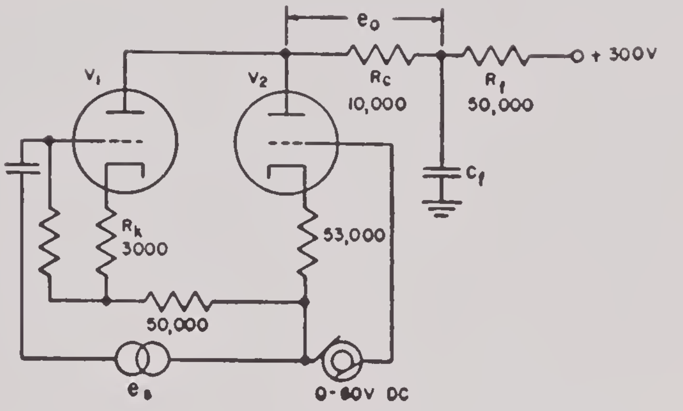


Figure 6. Circuit diagram of wave generator and cathode-ray indicator.

29.3.2 Monitor Oscilloscope

Throughout the design of the monitor oscilloscope, conventional television technique has been followed. All amplifiers have been carefully compensated for low frequencies to insure accurate reproduction of the complex wave. Linear tubes were selected and their characteristics further improved by adequate use of degeneration.



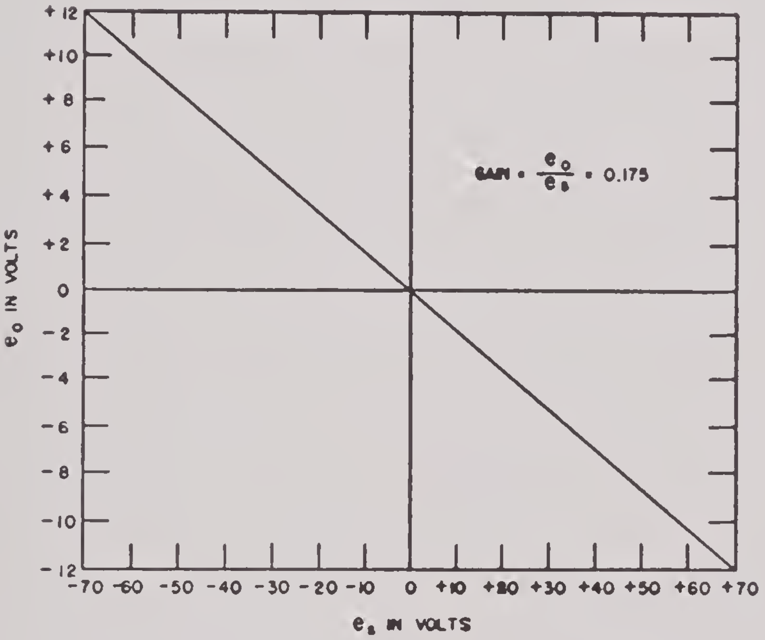


FIGURE 7. Operation of 6SN7 as mixer showing linearity.

The sawtooth voltage used for horizontal oscilloscope deflection and for synchronizing external equipment is generated in a very simple manner. Capacitor C_{17} (Figure 6) is charged from the 300-volt supply through resistance R_{69} and periodically discharged through commutator rings No. 2 and No. 4. The time during which the brush is passing over segments 41 and 42 has been assigned for synchronizing purposes. Consideration of Figure 6 will show that appropriate segments are located on commutator ring No. 2 to achieve this discharge in the allotted time. It will also be noted that C_{17} is prevented from recharging until the start of the first rectangular pulse. This is necessary since, if C_{17} was permitted to recharge immediately, a part of the synchronizing time would appear on the oscilloscope screen. The waveform of the voltage developed across C_{17} is plotted in Figure 8.

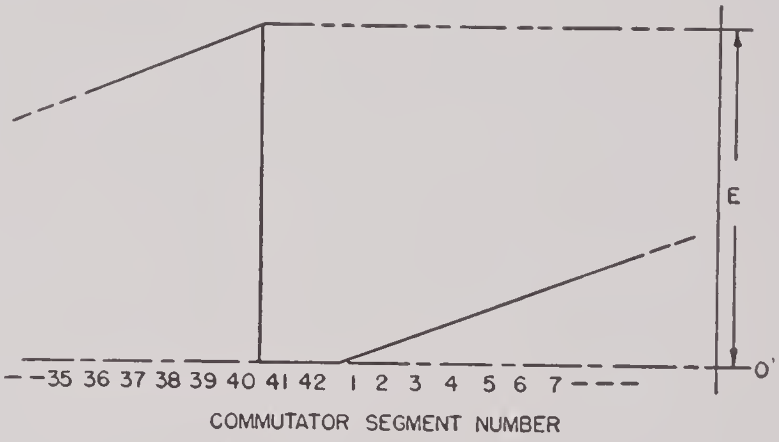


FIGURE 8. Sawtooth voltage wave developed across C_{17} charging through R_{69} from the 300-v supply and discharging through the commutator segments.

29.3.3 Alarm and Indicating Arrangement

In a push-pull oscilloscope deflection arrangement, one of the deflection plates must go positive with respect to the centering potential for either an upward or a downward motion of the oscilloscope trace. Consequently, by direct coupling the grids of a pair of thyratrons to the vertical plates of the oscilloscope, one thyratron can be biased to ignite on an upward deflection of the electron beam and the other on a downward deflection.

The flow of plate current that is initiated when either or both of the thyratrons Th_1 and Th_2 are ignited energizes a relay as shown in Figure 6. This relay, in turn, can be used to operate any type of alarm desired. Since the grid of a thyratron loses control over the plate current after the tube is conducting, the relay will remain in its alarm position until an attendant opens either switch SW_3 or SW_4 (see Figure 6). This action interrupts the flow of plate current through the thyratrons, thus de-energizing the relay.

To provide an indication of the direction of an unbalance between the two complex waves, an inherent characteristic of the thyratrons has been employed. When the thyratron is ignited, its grid endeavors to assume a potential somewhere between the potentials of its cathode and plate. Consequently, since the impedance of the deflection circuit is high, the conducting thyratron grid changes the d-c level of the particular deflection plate to which it is connected. This action changes the centering of the oscilloscope in the direction of the original unbalance between the two complex waves. The shift in oscilloscope centering, like the relay closure, remains until an attendant opens either switch SW_3 or SW_4.

29.4 APPLICATIONS OF THE SYSTEM

As the development work advanced, it became evident that modified forms of the basic cancellation system could be used in several applications other than those originally proposed.

29.4.1 Electrical Cancellation

This application utilizes the four functional units shown in Figure 9. It is assumed that an unknown wave of repetition frequency equal to the fundamental frequency of the complex-wave generator is available in current or voltage form and that it is desired to automatically excite an alarm system when changes occur in the unknown wave.

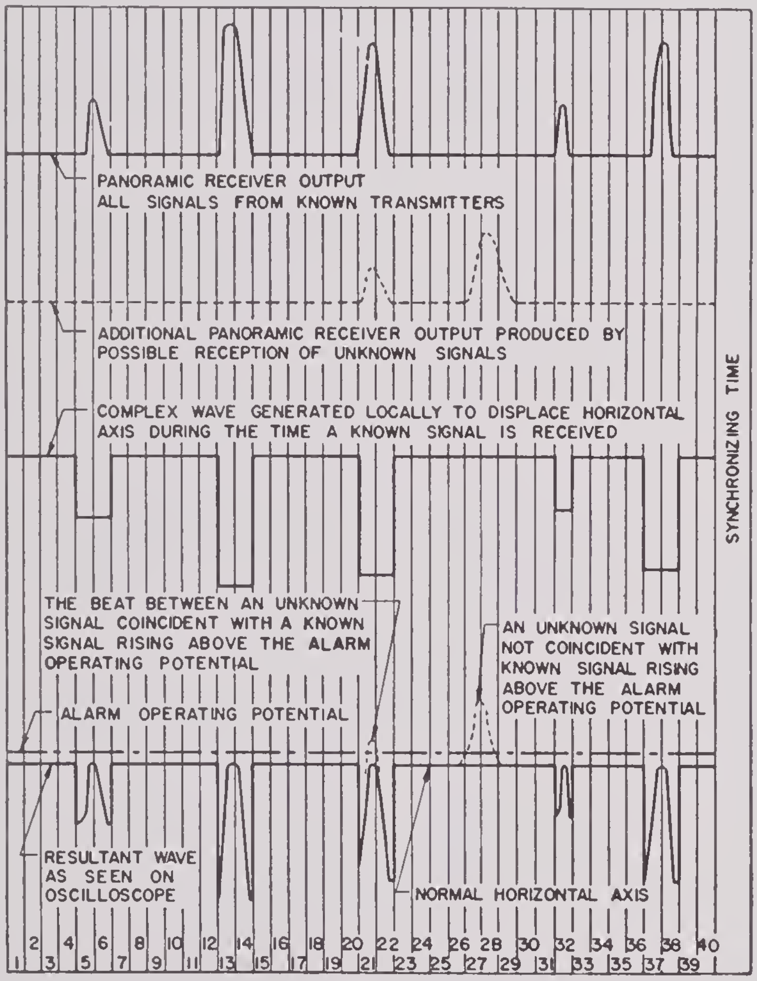

PANORAMIC RECEIVER OUTPUT
ALL SIGNALS FROM KNOWN TRANSMITTERS
ADDITIONAL PANORAMIC RECEIVER OUTPUT PRODUCED BY POSSIBLE RECEPTION OF UNKNOWN SIGNALS
COMPLEX WAVE GENERATED LOCALLY TO DISPLACE HORIZONTAL AXIS DURING THE TIME A KNOWN SIGNAL IS RECEIVED
SYNCHRONIZING TIME
THE BEAT BETWEEN AN UNKNOWN SIGNAL COINCIDENT WITH A KNOWN SIGNAL RISING ABOVE THE ALARM OPERATING POTENTIAL
AN UNKNOWN SIGNAL NOT COINCIDENT WITH KNOWN SIGNAL RISING ABOVE THE ALARM OPERATING POTENTIAL
ALARM OPERATING POTENTIAL
RESULTANT WAVE AS SEEN ON OSCILLOSCOPE
NORMAL HORIZONTAL AXIS
1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20 21 22 23 24 25 26 27 28 29 30 31 32 33 34 35 36 37 38 39 40

Figure 9. Drawing of possible cathode-ray screen pattern showing how alarm might be given when wave not cancelled by apparatus described makes its appearance.

The known wave from the commutator assembly and the unknown wave are combined in phase opposition at the mixer as previously described. A monitor oscilloscope at the mixer output then provides a sensitive measurement of the degree of approximation obtained after the potentiometers and their combined shunt capacitance have been properly set.

The alarm system in this case is essentially a tube-operated relay which, when energized, closes a power switch to an ordinary doorbell. By suitably biasing the tube below cutoff, the mixer-output noise residual can be rendered ineffective insofar as the tube plate current is concerned. An alarm threshold is thus established. If a change occurs in the unknown wave *regardless of polarity,* the mixer output will *increase* in direct correspondence. The manual setting of the alarm threshold determines the amount of change required to operate the alarm.

A visual illustration of the operation is best obtained on the oscilloscope. For example, with no local wave, the unknown wave only will show on the face of the tube. As potentiometers 1 to 40 are progressively adjusted, the unknown wave is sliced away until only a small noise residual remains. Suppose now we draw a hypothetical line just above the noise. This line corresponds to the alarm threshold. As an alternative, we might consider the average value of the noise to be the alarm threshold and then displace the noise downward. In either case, the effect is the same.

29.4.2 Panoramic Reception

The electrical cancellation and indicating system can be ideally applied to panoramic reception. The functional units of Figure 5 also provide a sawtooth wave of the same fundamental frequency as the local wave. If this wave is used to sweep the heterodyne oscillator frequency of the panoramic receiver, the output of this receiver can be cancelled out in identically the same manner as described above. Actually, if the panoramic receiver sweeps over a relatively wide spectrum with respect to its effective i-f pass band, the receiver output will consist of short pulses, each corresponding to a received r-f carrier. These pulses may require only a few segments from the local generator to displace them below the alarm threshold.

One advantage of this method is that on and off keying of known carriers does not interfere with the normal cancellation principle. This is evident, since proper adjustment of the alarm threshold when the unknown carrier is on will merely leave a downward displacement in the line when the unknown carrier goes off. In the same way, fading of the known signal does not actuate the alarm. A further advantage of this method is that no portion of the received frequency spectrum is blocked off. For example, if a known carrier is displaced downward and an unknown carrier of the same frequency starts up, the two will beat together giving a received signal of maximum amplitude equal to the sum of the two signals. This peak amplitude will exceed the alarm threshold, in the practical case, so long as the amplitude of the second received carrier is greater than the upward displacement of the alarm threshold with respect to the normal horizontal axis.

It is thus seen that the cancellation and indicating system when used in conjunction with a panoramic adapter or when constructed as an integral part of a panoramic receiver, provides completely automatic operation after the existing known signals have been cancelled out. One might therefore expect that several such receiver systems could be monitored by one observer without severe optical strain, since he would use the oscilloscope only during the initial setting-up procedure and during those times when the alarm indicated the appearance of a foreign signal.

29.4.3 Coding Signal Generator

The complex-wave generator portion of Figure 5 can duplicate any wave shape of fixed repetition period. It is, therefore, an ideal coding signal generator. In a practical application, the 40 amplitude-control potentiometers might be ten-step voltage dividers numbered one to ten. The code designation for a particular wave would then consist of 40 consecutive numbers indicating the individual settings of the 40 voltage dividers. The total number of possible combinations, that is, the variety of coding waves available, is given by the relation

$$C_n = \frac{400!}{10!\,390!},$$

or approximately

$$C_n \doteq 10^{19}.$$

These combinations can be further complicated by adding various shunting capacitances across the output of the complex-wave generator.

A practical adaptation of this coding device for two-way operation is indicated in Figure 10. For example, let $F_c(t)$ be the coding wave and $F_s(t)$ be the signal intelligence to be transmitted. These waves may be scrambled in any desired manner in their common mixer circuit. If we assume the mixer functions as a mathematical multiplier, the coded wave will be $F_c(t) \times F_s(t)$. This wave arrives at the receiver where an identical coding generator has its output inverted and multiplied into the incoming wave giving $F_c(t) \times F_s(t) \times 1/F_c(t) = F_s(t)$. It will be understood that the complex-wave generators at each end must be running in time and phase synchronism. The reverse transmission operates in exactly the same manner. However, if the coding wave at the original transmission point is to be satisfactory for decoding as well, it cannot lead the incoming scram-

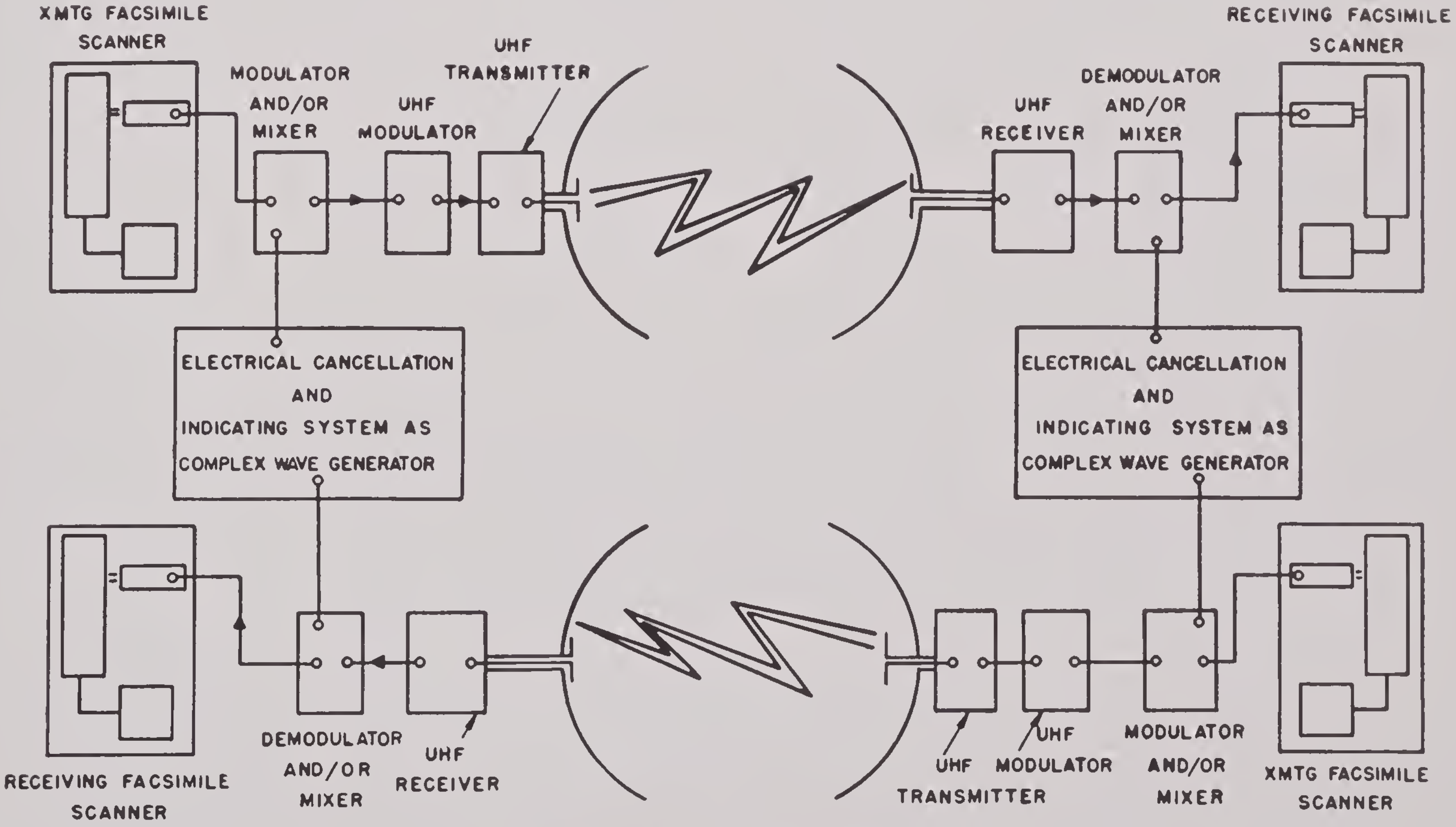


FIGURE 10. Use of equipment as coding signal generator for radio circuit.

bled signal by more than a few degrees. This means the transmission time around the radio loop must be small compared to the period of the coding wave. The arrangement of Figure 10 would therefore be useful for short-distance circuits only. Long-distance two-way operation could utilize an additional segmented ring and brush arm on the transmitting commutator. Proper phase adjustment could then be obtained by angular displacement of the brush arm.

29.4.4 Electrical Network Analysis

The introduction of radar equipment into the Military Services has necessitated rather extensive training programs for the operating and maintenance staffs. In particular, it is necessary that the personnel obtain a good knowledge of electrical network analysis, both steady-state and transient, in the shortest possible time. The instructor can either teach the required mathematical background including differential equations, the Fourier series, and the Laplacian transform as applied to networks, or teach the fundamentals of network theory and apply them heuristically to deduce the output response for applied force functions. The former method is essentially equivalent to the requirements for an engineering degree and little can be done to short-circuit the present three-year minimum time required. In this connection, it is believed that a modification of the complex-wave generator can serve as a very useful tool.

Suppose we incorporate in one unit a complex-wave generator, two oscilloscopes, and an assortment of ladder networks. The wave generator can furnish a large variety of waveforms to the inputs of the several networks and be observed on one of the oscilloscopes. Similarly, the output response of the networks can be studied individually on the second oscilloscope by appropriate switching. For classroom instruction, the oscilloscope faces should have at least 9-in. diameters and might include transparent scales to form a rectangular coordinate system.

A device of this type can be used to verify existing mathematical solutions of network responses. In addition, it can derive approximate responses of very complex networks for which no exact solutions have as yet been obtained. Moreover, it would appear that such a device would be a very useful tool for classroom instruction, particularly when it is necessary to reduce the instruction time to an absolute minimum.

29.4.5 Multiple Station Monitoring

There are certain minor applications of the device that might have merit in specific cases. If it is desired to monitor the *frequency* of several transmitters simultaneously, one might use the arrangement described above under Panoramic Reception. If the frequency of one carrier varied, it would be displaced on the time axis of the panoramic scope and rise out of its declivity, thus ringing the alarm. If it is desired to monitor the *transmissions* of several transmitters simultaneously, one might have several receivers with their outputs connected to the individual segments of a commutator. The rotating brush arm would then serve as a scanner with its output connected to a page-type facsimile recorder. By synchronizing the drum speed of the recorder with the rotation of the brush arm, each transmitter signal would be recorded as a horizontal track similar to variable-density movie records.

Chapter 30

LAYING FIELD WIRE BY AIRPLANE

Laboratory investigation and field tests leading to methods of packing field wire in a suitable form for paying out from an airplane in flight, and to techniques for laying the wire at speeds up to 200 mph.

30.1 HISTORY OF THE DEVELOPMENT

The interest of NDRC in air-laid wire originated in a request for a study of listening devices to detect enemy movements in jungle areas. It was suggested to NDRC in July and August 1942 that it might be possible to carry listening devices over into enemy territory by planes. It was thought that planes could string numerous long camouflaged wire sections over treetops and ahead of advance positions. Small microphones at the wire ends would enable listeners to detect the position and extent of enemy movement. It was also believed that the ends could be laid with sufficient accuracy to permit sound ranging on gun positions.

This early interest later extended to possibilities of laying wire from planes for special communication purposes, with emphasis largely on jungle terrain and with the hope of materially reducing the hazard of direction finding, jamming, and interception.

Accordingly Project C-72[a] was initiated to investigate the possibilities of laying wire by air.

After a preliminary survey the problem divided itself into several lines of investigation: selection of the wire, behavior of the wire during uncoiling and passage to the ground, determination of the best form of package for the wire, technique of laying wire, and winding machines. The final report[1] on Project C-72 gives the results of these studies.

After considerable laboratory work, a few trial flights were made at Fort Knox in April 1943. The results were so encouraging that the Army gave wide circulation to the NDRC report. This report eventually reached the China-Burma-India theater where conventional methods of laying wire were not only inadequate but fraught with danger. In January 1944, Army Air Forces authorities in the CBI theater cabled a request for a continuation of the development work which had stopped because of termination of the project. Shortly after this the Air Technical Service Command [ATSC] requested that the work be resumed on a rush basis, and a contract covering the completion of the work was negotiated directly between the Western Electric Company, Inc., and ATSC.

30.2 RESULTS OBTAINED

The following summary includes the work accomplished under the NDRC contracts as well as the direct contract mentioned above.

An outdoor laboratory was set up at Murray Hill, New Jersey, including a machine capable of "pulling" wire from packs and coils at speeds up to 200 mph. Many fundamental tests were made at this location and these were followed by more than 200 flight tests at the Army Air Base, Fort Dix, New Jersey.

Three methods of packing wire were investigated: the carpet pack, the solenoid coil, and the crisscross coil equipped with center guides. The last system proved most practicable and was eventually standardized. This system was found capable of laying WD-1 wire in 16-mile lengths at speeds up to and exceeding 150 mph. It was found possible to lay Canadian D-8 wire and standard W-110B in 16-mile lengths at somewhat slower speeds. Following the experimental flights at Fort Dix, New Jersey, a demonstration run was made for the Army with complete success over a rugged 16-mile course in the Great Smoky Mountains.

Ten machines capable of winding field wire in crisscross coils were assembled and turned over to the Air Forces. Also, five complete equipments for laying wire from airplanes were furnished. Included with the latter was a manual covering the entire technique of winding and laying wire at high speeds.

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[a] Project C-72, Contract OEMsr-879, Western Electric Co., Inc.

Chapter 31

FLOATING INSULATED WIRE

A lightweight wire of 700-yd maximum length was desired which would have sufficient buoyancy to float and be strong enough to pull a rubber raft.[a] In preliminary experiments made in the theater, a wire taped to a ½-in. rope was used. Coral cut the rope and there was a limit to the length that could be used because of the weight.

31.1 SUMMARY OF RESULTS

Since the specific gravity of polyethylene (0.92) is sufficiently below that of sea water to float, it was thought that a floating wire of considerable strength could be obtained with reasonable dimensions. The tensile strength of polyethylene is relatively low, about 1,000 lb per sq in. for an elongation of about 25 per cent. Accordingly, an efficient design of floating wire would be one in which the polyethylene serves to provide buoyancy and insulation and the metallic conductor in the core provides strength.

The simplest design to produce is one consisting of a single conductor of stranded steel encased in the insulating material. This assumes a sea-water return for the electric circuit.

A smaller wire of equal strength could be provided by using Fiberglas continuous-filament tying cords for strength and a small copper wire for conductivity. The tensile strength of Fiberglas cords is approximately 75,000 to 100,000 lb per sq in. with an elongation that is low. The specific gravity is 1.5. Table 1 below gives a comparison of the two proposed methods of solving the problem.

Table 1. Comparison of steel core and Fiberglas-copper core floating wires.

Breaking strength (lb)	Outside diameter (in.)		Weight per 1,000 ft (lb)	
	Steel core	Fiberglas-copper core	Steel core	Fiberglas-copper core
300	0.31	0.23	33	18
500	0.40	0.27	56	25
1,000	0.57	0.36	111	44
2,000	0.81	0.49	222	82

Other materials, such as nylon and Fortisan, were considered and their possibilities for this application discussed.

Provisions for protecting the wire against coral reefs, the effects of ocean currents, and handling properties are discussed in the final report.[1]

31.2 CONCLUSIONS

The most promising design would employ a core of Fiberglas or other lightweight high tensile strength material and one or more small copper conductors and polyethylene insulation of sufficient thickness to float the wire. A size providing a tensile strength of 500 lb would have a diameter slightly over ¼ in. and a weight of about 25 lb per 1,000 ft. Greater strength could be provided by using a twisted pair of such wires.

The suggested design was speculative. It was expected that a year would be required to obtain wire in quantity for testing under field conditions, readying for production, and so forth. The matter was referred to the Services for such direct action as they might wish to make.

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[a] Project 13-104; Contract OEMsr-1413, Western Electric Co., Inc. Several matters were considered under this project as noted in the bibliography. Aside from the subject of this summary, the other matters were largely discussions of problems. Division 13 undertook no actual work on these subjects.

Chapter 32

NOISE INVESTIGATIONS

Several projects under Division 13 were concerned with the problem of noise from power-generating machinery. The question of audible noise from mobile field generators was covered by an investigation of methods of quieting such apparatus conducted as part of the broad work carried out in Project C-79.[1] Two other projects, one dealing with noise measuring equipment, are being continued by the Services after the NDRC contracts were cancelled in October 1945.

32.1 NOISE IN AIRCRAFT ELECTRICAL MACHINERY

Under this project[a] experimental studies were to be conducted to determine the factors affecting the generation of r-f voltage in 28-volt d-c aircraft rotating electric machinery and to obtain information leading to the design of such machinery so that the radio noise produced would be minimized and would lead to the simplification of filtering such machinery.

During the period of the project,[2] 30 pieces of equipment obtained from Wright Field and from the Bureau of Aeronautics were examined. They included motors, dynamotors, inverters, etc., and the measurements included conducted noise and noise field intensity.[3]

At the end of the project, the sample equipment had been tested and charts prepared (included in the terminal report) to show the noise levels measured. Analysis had been carried to the point where some of the features associated with radio noise were evaluated and procedures for investigating other features were indicated.

In the tests, conducted noise on the 28-volt circuit ranged from 1 to 12,000 μv. Some of the samples had capacitors or filters. The "50 per cent curve," plotted from the noise values that were equalled or exceeded in 50 per cent of the tests at the respective frequencies, ranged from 25 to 750 μv, a ratio of 1 to 30.

Conducted noise on the output of dynamotors with resistance load varied directly as the machine voltage and inversely as the frequency at which the noise was measured. In different machines at the same voltage, the noise varied inversely as the number of commutator bars between brushes.

The conducted noise in μv on the dynamotor output circuit was approximately equal to

$$\frac{1300 \times \text{d-c volts}}{(\text{Bars per brush})\,(\text{Frequency in mc})}\,.$$

The results pointed to the fact that most of the radio noise produced by the samples was caused by the passage of the brushes from bar to bar of the commutator.

32.2 NOISE MEASUREMENT

The subject of this project[b] was "to conduct miscellaneous studies and investigations involving radio interference phenomena including development of standards for measurements and measurement techniques, comparison of various types of noise meters and construction of a standard noise meter." This work will be carried forward under contract with the Bureau of Ships.

The need for investigation of radio noise measurements arises from the fact that such measurements using noise meters now available show anomalous discrepancies. It is reported that various instruments having been calibrated with sine wave signals, read differently to as great an extent as 1,000 to 1 when they are presumably the field intensity from the same noise source. With the same meter very different results are frequently obtained when readings are taken on two frequency bands in the region of their overlap. The degree of discrepancy is reported to be different for different noise sources. The discrepancy is greater when radiated noise is being observed than when direct conducted noise is being observed.

Under the project, representative noise meters were to be examined critically in a screened room. At the termination of the contract, the noise meters had not been received from the manufacturers, so that the experimental program was not begun. Certain topics were taken up, however, including a search of the existing literature, examination of reports concerning the performance of existing noise meters, and certain mathematical analyses which will be useful when the experimental phase of the investigation is possible.

The final report[4] contains the analysis of present noise meters, a definition of noise, present noise-meter specifications as set up by the Joint Coordination Committee on Radio Reception of the Edison Electric Institute, the National Electrical Manufacturers Association, and the Radio Manufacturers Association. Methods of coupling a noise meter to a source are considered. The appendices contain discussions of antennas for noise meters, detectors of electrical noise, methods of calibrating a noise meter, the problem of frequency conversion, and a bibliography.[1-4]

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[a] Project 13-117, Contract No. OEMsr-1475, General Electric Co.

[b] Project 13-123, Contract OEMsr-1478, University of Pennsylvania.

INTERFERENCE REDUCTION COMMITTEE

Until late in World War II, the problems of electrical interference between pieces of electronic equipment produced by the several NDRC divisions had not reached the point where the situation was critical, although much good could have been done had some organization existed from the start to minimize such interference. Typical examples of such difficulty were the interference in communication equipment created by certain types of radar and the spurious responses registered on countermeasure search receivers caused by other electronic apparatus mounted in the airplane carrying the search receiver.

The effectiveness of only a small amount of coordinating activity was fully appreciated in October 1944 when an informal meeting was held at which representatives of the Army, the Navy, and NDRC discussed the organization of such a group.

In January 1945, the Interference Reduction Committee [IRCOM] was established by Ward F. Davidson as a special committee of the Chairman's Office of NDRC, with Alan Hazeltine as chairman and with representatives from Divisions 5, 13, 14, 15, and 17 of NDRC. Although the life of the committee was short, terminating in October 1945, certain accomplishments were made, and enough experience gained to indicate the importance of such a coordinating agency and to set the pattern for future work in the subject.

IRCOM Accomplishments

To limit its activities and thus to work most effectively, the committee excluded from its interest all interference caused by "administrative decision," for example, if in a given campaign a radar and a communications system interfered with each other by being on the same frequency, administrative action would be required to shift one of the frequencies or to take one of the Services off the air. Enemy jamming was beyond the scope of the committee, as were precipitation static and meteorological phenomena causing interference. Interference problems which fell entirely within the sphere of a single NDRC division were similarly excluded.

IRCOM, therefore, considered matters where interference was caused accidentally by electronic devices which, given a sufficient refinement in design, should be able to work together in harmony. By and large, its work was mostly concerned with interference to communication services by noncommunication systems.

Standards of Measurement

The principal difficulty with an intelligent study of the interference problem proved to be the lack of standardized methods of making interference measurements. The Army and Navy, independently and on their own initiative, were attempting to establish specifications to standardize such methods of measurement so that limits on interference could be written into new contracts for electronic equipment. IRCOM took on this problem of measurements and produced a report on noise generators and noise meters.

One of the principal jobs of IRCOM was to consider potential NDRC contracts referred to it. Its recommendations on interference measurements were as follows:

Using the 931 vacuum tube as a source of "white" noise, the development of a stable noise generator should be undertaken by one of the Services.

A study of the stability of the Detroit Signal Laboratory noise generator should be carried out at the Psycho-Acoustic Laboratory, Harvard University.

When the hydrogen thyratron tube reaches the proper point of development, a pulse generator should be constructed around it as the nucleus.

Exhaustive tests should be made of the noise generator being developed by Purdue University (Project 13-113, AC-238.04, OEMsr-1431) to further development work along similar lines or to aid in production of models.[5-6]

Chapter 33

TRANSIENT RESPONSE OF BAND-PASS AMPLIFIERS

Analysis of band-pass network response, including i-f, r-f, and stagger-tuned amplifier response to delta functions, i.e., pulses of zero duration, infinite amplitude, and finite area.

33.1 INTRODUCTION

The response of a band-pass amplifier[a] to signals with waveforms having severe discontinuities is of great interest for many applications. For example, it is common practice to put a limiting circuit in a radio receiver to reduce interference from static. Such a limiter allows a signal to pass through the audio section of the receiver only if its amplitude is less than some predetermined maximum. Noise of amplitude greater than this is clipped and its effect greatly reduced. If the limiter is to remove extraneous noise effectively, that noise must be present in the form of pulses of large amplitude and short duration at the input to the limiter. Static coming into the receiver often approximates a delta function (a pulse of zero duration and infinite amplitude but having finite area) in waveform. The r-f and i-f amplifiers of the receiver constitute a band-pass network.

Equations derived and given in the final report[1] indicate that to obtain output pulses of short duration and large amplitude, the frequency-response curve of r-f and i-f amplifiers should have a large band width with sides which slope very gradually. Of these two requirements the latter is more important. If the band width is increased, the primary effect is to increase the amplitude of the pulse. This results because the increased band width allows the circuits to obtain more energy from the input delta function. The large amplitude pulse is easily removed by a limiter, but there is a greater amount of energy which must be removed. Since a limiter clips at a fixed rather than a relative voltage, the duration of the truncated pulse passing through the limiter will be increased if the geometrical shape of the pulse is maintained and its amplitude increased. If the slope of the side of the response curve is decreased, maintaining constant band width, the amplitude of the pulse remains almost constant. The shape of the pulse is changed so that the truncated pulse from the limiter is of shorter duration.

The calculation of the delta-function response for a complex amplifier network is unfortunately a tedious problem. In a Telecommunications Research Establishment report, Twiss has calculated the step-function response of a number of networks of the type described by Wallman[2] as stagger-tuned circuits. In these calculations he has replaced the actual band-pass r-f network with its video analogue and so obtains directly the envelope of the response. The band-pass network having a steady-state response curve symmetrical about its mid-frequency is replaced by a hypothetical low-pass network having a response curve identical except that it is symmetrical about zero frequency. The transient response of this low-pass network is directly the envelope of the response of the band-pass network. This procedure is valid only provided the band width is small compared with the mid-frequency. The fact that it yields the envelope of the response is usually a distinct advantage, since the envelope is normally the feature of greatest interest.

For such stagger-tuned networks Twiss[3] has shown that the steady-state response as a function of frequency may be expressed as

$$g = \frac{1}{(1 + x^{2r})^{n/2}}\,, \tag{1}$$

where g is the relative response;
r is the number of tuned circuits in one staggered unit (one r-uple);
n is the number of such units;
x is the normalized frequency variable $2f/B_1$;
f is the frequency measured from the mid-band frequency;
B_1 is the band width of one circuit, measured between frequencies where the response is $1/\sqrt{2}$ times the maximum.

This equation assumes that the band width of the network is small compared with its mid-frequency. Twiss gives calculations for networks of this sort for a reasonably wide range of values of both r and n. He further points out just what physical circuits correspond to these mathematical values. His data are given in the form of curves plotting the voltage output due to a unit step-function input as a function of the normalized time variable, $(t/t_0) - \pi B_1 t$.

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[a] Project 13-110, Problem No. 4, Contract No. OEMsr-1441, Harvard University, in collaboration with Project NS-108 of Division 17.3 NDRC, Psycho-Acoustic Laboratory, Harvard University.

In the present study the primary interest was in obtaining some sort of approximate relations from which important characteristics of the delta-function response could easily be calculated. If possible, these relations should apply to all band-pass systems in general and should be simple and accurate enough for engineering use. The procedure used was to find the delta-function response of the stagger-tuned networks of Twiss by finding the time derivative of his step-function response curves, making use of a curve-fitting process. These delta-function responses were then related to the shapes of the steady-state responses of the networks.

33.2 RESULTS OF ANALYSIS

The response of a stagger-tuned band-pass network to a delta function may be approximated by two empirical relations. The first of these is

$$\begin{aligned} t' &= 2.6\,\phi = 0.21\,\Phi \\ t'' &= 1.8\,\phi = 0.15\,\Phi \end{aligned} \tag{2}$$

where t' is the duration of the delta-function response measured 60 db below its peak, and t'' is the duration measured 40 db below the peak. The slope of the side of the steady-state frequency-response curve is ϕ, where g is the relative response and the slope is measured at $g = 1/\sqrt{2}$. If the curve is expressed in decibels and the slope is measured at $G = -3$ db, the slope is Φ. In either case the slope is measured in terms of a frequency scale in units of cycles per second.

The second relation is

$$e_\delta = B_n + \frac{0.6}{\phi} = B_n + \frac{7.5}{\phi}, \tag{3}$$

where e_δ is the peak of the delta-function response and B_n is the overall band width of the network measured at $g = 1/\sqrt{2}$, or at $G = -3$ db. This peak is for an amplifier network of unit voltage amplification and for a delta function of unit area.

These relations were obtained empirically for certain stagger-tuned networks, but it is contended that they are equally useful for band-pass networks in general, so long as the band width of the network is small compared with its mid-frequency.

33.2.1 Typical Example

As an example of the application of the approximate relations, consider a typical i-f amplifier with the following characteristics:

Mid-frequency, 465 kc.
Band width, 3 db down, ±5 kc.
Slope of response curve, 3 db down, 2.5 db/kc.
Voltage amplification, 1,000.
Input-pulse duration, 0.5 μsec.
Input-pulse amplitude, 10 volts.

Since the input-pulse duration is much smaller than the reciprocal of the mid-frequency, the equations derived may be used to find the duration of the pulse at the output of the amplifier and its amplitude. The duration is

$$t' = 0.21\,\Phi = 0.21 \times 2.5 \times 10^{-3} = 525\ \mu\text{sec.}$$

$$t'' = 0.15\,\Phi = 0.15 \times 2.5 \times 10^{-3} = 375\ \mu\text{sec.}$$

Assuming that the input pulse is rectangular, its area a is

$$a = 10 \times 0.5 \times 10^{-6} = 5 \times 10^{-6} \text{ volt seconds.}$$

The amplitude of the output pulse is

$$e_\delta = B_n + \frac{7.5}{\Phi} = (5 + s) \times 10^3 + \frac{7.5}{2.5 \times 10^{-3}}$$
$$= 13{,}000 \text{ volts.}$$

This amplitude is for an input pulse of unit area and does not consider the voltage amplification of the amplifier. Taking account of these factors, the actual amplitude of the output pulse is

$$\begin{aligned} e_\delta &= \text{area} \times \text{voltage amplification} \times 13{,}000, \\ &= 5 \times 10^{-6} \times 1{,}000 \times 13{,}000, \\ &= 65 \text{ volts.} \end{aligned}$$

33.3 CONCLUSIONS

While the approximate relations for the calculation of the duration and peak of the output response of a network which has a delta function applied to its input are admittedly empirical, they give results which are of engineering value for the stagger-tuned networks from which they were derived. Of course, this gives no assurance that they are equally valid in the case of any other arbitrary network. They are of the correct form to agree with experimental observations, however, and it is felt that they should be useful in the design of any band-pass system whose band width is small compared with its mid-frequency. Qualitatively they may lead to useful generalizations.

Chapter 34

MEASUREMENTS OF MAGNETIC PROPERTIES OF FERRITE CORE MATERIALS

Examination[a] of the characteristics of certain ferrite materials brought to the United States from the Phillips Company, Eindhoven, Holland, in May 1945. Two of the materials, MnZn ferrite, appeared to be suitable for low frequencies of 10 to 60 kc, and the third, CuZn ferrite, was less suitable for low frequencies but showed possibilities at frequencies as high as several megacycles. All were unstable with regard to temperature and voltage.

34.1 MEASUREMENTS

THE CORE SAMPLES were in toroid form and it was only necessary to wind upon them an appropriate number of turns (5 to 215) for a given frequency range and to measure values of inductance and either Q or the apparent resistance. The relations of the several coil constants are as follows:

Magnetic Permeability, μ. For a closely wound toroid on a uniform core, the permeability and the inductance are related by

$$\mu = \frac{lL}{1.256N^2A \times 10^{-8}}, \tag{1}$$

where l is the effective length of the flux path in cm, L is the inductance of the coil, N is the number of turns and A is the cross section area of the core in sq cm.

Specific Loss Factor, ρ. The power dissipated in the coil carrying an alternating current is

$$W = I^2R, \tag{2}$$

where R is the total equivalent resistance of the coil. This resistance can be divided into R_c, the resistance of the coil without the core, and R_i, the resistance produced by the core losses. If W_i is the power dissipated because of the core then

$$W_i = I^2R_i. \tag{3}$$

Now ρ is defined as

$$\rho = \frac{W_i}{B^2V} = \frac{I^2R_i}{B^2V} \frac{\text{watts/cm}^3}{\text{gauss}^2}, \tag{4}$$

where V is the volume of the core material and B is the magnetic induction.

For a toroid

$$B = \frac{4\pi NI\mu}{10\,l} \text{gausses}, \tag{5}$$

where l is the length of the flux path. Then

$$\rho = \frac{100I^2R_il^2}{16\pi^2N^2I^2\mu^2V}. \tag{6}$$

The inductance L is

$$L = \frac{4\pi N^2I\mu A}{10lI \times 10^8}, \tag{7}$$

and, since $V = lA$,

$$\rho = \frac{R_i}{4\pi\mu L \times 10^7} \frac{\text{watts/cm}^3}{\text{gauss}^2}. \tag{8}$$

Values of R_i were obtained by measuring the total resistance and subtracting from it the resistance of a similar coil without the core. At frequencies below 100 kc the d-c resistance was practically the same as R_c.

Q Measurements. R was obtained by means of a Q meter which measured the ratio of the inductive reactance $L\omega$ to the total equivalent resistance R of the coil and core. The Q meter gave values of Q and of C, the capacitance required to tune the coil to a given frequency. Plotting $1/\omega^2$ as a function of C gave a graph whose slope is L, the true inductance. From this value and that of Q, the resistance can be obtained. Thus

$$R = \frac{L\omega}{Q}.$$

By subtracting R_c, the resistance of an identical coil with an air core, R_i is determined and can be calculated by equation (8).

34.1.1 Equipment Employed

In the range between 60 and 400 kc, a Boonton Type 100-A Q meter was used with a precision capacitor (General Radio) in parallel with the Q meter capacitor for greater precision.

In the range of 400 to 5,000 kc, a General Radio Type 916-A r-f bridge was used. Trouble was encountered in the range 1 to 140 kc because of the necessity of applying too much voltage to the coil. By using a General Radio Type P-508 bridge as Owen bridge for measuring inductance and as a resonant bridge for measuring Q, ρ could be calculated.

With the Owen bridge and the several resonance bridges employed, balance was indicated by an oscilloscope preceded by a tuned circuit and 200 db of available amplification. Visual discrimination between the fundamental and harmonic aided in detecting correct balance.

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[a] Project 13-110, Problem No. 8; Contract OEMsr-1441; Harvard University.

34.2 MAGNETIC PERMEABILITY

The magnetic permeability μ was virtually constant over the frequency range used. Some variations were apparently due to changes in temperature, to changes in applied magnetic intensities, and to flux leakage. Measured values are as follows:

Material	μ (avg)	Range of μ
MnZn ferrite A	757	734 - 770
MnZn ferrite B	846	834 - 860
CuZn ferrite	392	382 - 396

The final report[1] gives all the data taken and shows that the materials were superior to molybdenum permalloy in that they have higher permeabilities and lower losses but are inferior in stability. A marked change in properties occurred when the material was subjected to high magnetization — 1,000 gausses. A permanent change was caused, although for several days the permeability slowly changed in the direction of its former value.

Although the materials showed promise, it was concluded that they would be useless unless means were found to stabilize their characteristics to reduce both temperature coefficient for μ and ρ and to reduce the change in values which occurs when magnetic induction values greater than 10 gausses were applied.

Chapter 35

AIRCRAFT ANTENNA POWER, IMPEDANCE, AND TUNING-NETWORK SURVEY

Study to determine tuning and loading range of certain transmitters, methods of measuring antenna impedance, and methods for measuring antenna power.

35.1 PROJECT SCOPE

The objectives of this project[a] may be summarized as follows:

1. The development of a method and necessary equipment for determining the range of complex antenna impedances into which a given transmitter will tune the load properly, proper loading being defined as that which will draw normal plate current when the tuning control is set for minimum plate current, measurements to be made at the antenna terminals and/or at the power-amplifier plate terminals.

2. A survey of the ranges of complex antenna impedance into which specific aircraft transmitters will tune properly as defined above. The antennas into which these transmitters operate may have resistive components ranging from ½ to 15,000 ohms, and reactive components ranging from —15,000 to +15,000 ohms.

3. The development of a method and equipment for measuring the resistance and reactance of an aircraft antenna within an accuracy of 5 per cent and over a frequency range of 1 to 20 mc. The method is to be simple in application and is to require no extensive calculations.

4. The development of a method and test equipment for measuring the power output of a transmitter into its antenna over a range of 5 to 125 watts within an accuracy of 5 per cent over a substantial portion of the range.

At the end of the war and termination of the contract, no work had been accomplished on 4, which is being carried on under contract with the Office of Naval Research.

35.2 TRANSMITTER ANTENNA IMPEDANCE RANGE

This problem is not simple, largely because of the wide range of frequencies and of antenna impedances which must be considered, and also because of the variety of circuits in the power-amplifier stage of military transmitters.

Two methods have been tried for determining the loading range. In one method a dummy antenna or adjustable load is used for actually loading the transmitter. The other method obtains the loading range from impedance measurements made at the power amplifier and antenna terminals.

35.2.1 Method Using Adjustable Loading Unit

The adjustable load provides the combinations of resistance and reactance normally encountered in the operation of aircraft transmitters in the 2- to 18-mc range. In addition, it facilitates measurement of the transmitter power output and hence provides data giving the useful loading range of the transmitter. The variations of resistance and reactance are obtained through the use of several circuits composed of variable inductance, variable capacitance, and resistance. Physically large circuit elements are used to dissipate 125 watts or more and to withstand the large currents and voltages which may exist.

It was found impractical to provide an accurate calibration of the adjustable loading unit. The stray capacitances due to wiring, the many possible combinations of components, and effects due to heating, all preclude reliable calibration of the unit. While approximate calibration curves have been prepared to assist in the adjustment of the loading unit, a more accurate value of the loading impedance was obtained by measurement with impedance bridges.

The application of the adjustable loading unit for determining the loading range of a power transmitter is described in the final report.[1] Time permitted only a partial survey of the BC-375-E transmitter, but the loading range provided by each of its four antenna circuits was determined at one frequency, 4 mc.

35.2.2 T-Network Method

Section IV of the final report describes a method for determining the loading range of a transmitter by means of a series of impedance measurements. The method utilizes the facts that (1) a tuning and loading network can be represented by a T-network and (2) a properly tuned and loaded power-amplifier tube operates into a plate load which is a pure resistance R_L. By measuring the resistance R_L suitable for the power-amplifier tube, by making open-circuit impedance measurements at the power-amplifier (Z_{O1}) and antenna terminals (Z_{O2}), and by making a measurement at the power-amplifier terminals with the antenna terminals short-circuited (Z_{S_1}), the unknown (antenna) impedance Z_x can be computed from

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[a] Project 13-110, Problem No. 6, Contract No. OEMsr-1441, Harvard University.

$$Z_x = Z_{O2}\frac{R_L - Z_{S_1}}{Z_{O1} - R_L}.$$

The plate load R_L must be measured whenever the operation of the power-amplifier tube changes with frequency, and the other measurements must be made for a large number of adjustments of the tuning and loading controls. It should be observed that the extreme settings of the tuning and loading controls do not completely define the range of impedances which may be used with a given antenna circuit.

The T-network method has been used at a single frequency in determining the tuning and loading range provided by four of the thirteen circuits present in the AN/ART-13 transmitter.

35.2.3 Conclusions

In the somewhat limited application of the two methods, certain opinions were formed which may assist in their evaluation. These may be summarized as follows:

1. Methods using an adjustable loading unit and T-network measurements will each yield the results desired.

2. Both methods require the availability of impedance-measuring equipment, and one requires an adjustable loading unit in addition. In any event, facility in transmitter operation and in the use of r-f impedance-measuring equipment is required.

3. Both methods are time consuming because of the large amount of data required to specify completely a transmitter loading range. However, the method which utilizes an adjustable load is more direct and appears to have the advantage in this respect.

4. The relative accuracies of the methods depends upon the circuit being analyzed. The accuracy of the T-network method is reduced when its application requires measurement of the impedance of parallel resonant circuits. The results obtained with the adjustable load are more indefinite in those cases in which transmitter tuning and loading adjustments are broad.

5. The adjustable load facilitates measurements of power output; the T-network does not. This is an important consideration, for the impedance loading range as determined by rated plate current may not be the useful loading range as determined by r-f power output.

35.3 COMPLEX IMPEDANCES OF AIRCRAFT ANTENNAS

A survey has been made of the literature and of present-day impedance-measuring equipment applicable to this assignment. The bibliography included in the final report contains the most pertinent references. For such an important subject, surprisingly little work has been reported. Of the commercial instruments available, the Boonton Radio Corporation Q-Meter line and the General Radio Company companion pieces, the Type 916-A R-F Bridge and the Type 821-A Twin-T Impedance-Measuring Circuit, seem to be most applicable at the present time.

None of the available impedance-measuring equipment has been found to combine the requirement of simplicity of operation (without extensive calculations) with that of 5 per cent accuracy over the wide ranges of impedance and frequency. Furthermore, the impedance bridges with their required accessories are considered to be too bulky and heavy to be used conveniently in airplanes in flight. Accordingly, plans have been directed toward the modification of existing methods and equipment and toward the development of new methods.

Methods under investigation, which include certain original features, are the following: a substitution which utilizes a decade r-f resistor unit, a method which uses substitution (dummy) antenna units to provide duplicate values of antenna impedances so that measurements may be performed in the laboratory, and a low-power transmitter which utilizes power-amplifier plate current to measure both resistance and reactance. Time permitted only partial investigations of these methods. Pending the outcome of this work, no effort has been made to adapt commercial equipment for this purpose.

The following additional conclusions are indicated in connection with the impedance-measuring methods considered:

The possibility of using small dummy antenna units for transferring the antenna impedance characteristics to the laboratory seems entirely feasible. A more direct method may be preferred even though it be somewhat less accurate.

The principle of the impedance-measuring transmitter seems to satisfy the requirements of size, weight, and simplicity of operation. Incomplete tests indicate that the equipment may be made substantially direct-reading over a considerable portion of the impedance range.

BIBLIOGRAPHY

Numbers such as Div. 13-200.1-M1 indicate that the document listed has been microfilmed and that its title appears in the microfilm index printed in a separate volume. For access to the index volume and to the microfilm, consult the Army or Navy agency listed on the reverse of the half-title page.

Chapter 1

The final report on Project C-79 (Contract OEMsr-1018) was composed of three parts, each made up of numerous individual reports as indicated below:

1. *Systems Engineering for Army Air Forces Communications*, Part I, A. B. Clark, OSRD 1442, OEMsr-1018, Report 2519, BTL, Apr. 27, 1943. Div. 13-200.1-M1

    Report of Progress, A. B. Clark, Apr. 27, 1943.

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2. *Systems Engineering for Army Air Forces Communications*, Part II, A. Tradup, OSRD 1925, OEMsr-1018, BTL, Oct. 1, 1943. Div. 13-200.1-M2

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1. *Precipitation Static Reduction Research*, W. H. Huggins, OSRD 1907, OEMsr-92, Project C-21, Oregon State College, Mar. 31, 1943. Div. 13-202.1-M1

Chapter 8

1. *Precipitation Static Research,* Sheldon H. Dike, OSRD 1411, OEMsr-678, Project C-41, University of New Mexico, May 1, 1943. Div. 13-202.21-M1

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1. *Study of Effect of Aircraft Surface Treatment on Electrical Charges Causing Precipitation Static,* Roscoe H. George, OSRD 1446, OEMsr-679, Project C-64, Purdue Research Foundation, May 11, 1943. Div. 13-202.21-M2
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Chapter 12

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   Improved Panoramic Receiver, Part I, E. R. Taylor, OSRD 909, OEMsr-357, BTL, Sept. 1, 1942. Div. 13-203.3-M1
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Chapter 14

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   Improved Panoramic Receiver, Part I, E. R. Taylor, OSRD 909, OEMsr-357, BTL, Sept. 1, 1942. Div. 13-203.3-M1
   Improved Panoramic Receiver, Part II, E. R. Taylor, OSRD 909, OEMsr-357, BTL, Jan. 22, 1943. Div. 13-203.3-M2

Chapter 15

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Chapter 24

1. *Sound Recording on Magnetic Materials*, C. B. Jones and D. B. Davies, OSRD 3099, OEMsr-833, Project C6-69, Brush Development Company, Dec. 31, 1943. Div. 13-206.6-M2
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Chapter 25

1. *Possibilities of Substitute for Natural Quartz for Frequency Control at High and Ultra-High Frequencies*, OSRD 287, OEMsr-120, Project C-29, Brush Development Company, Dec. 6, 1941. Div. 13-207.1-M1

Chapter 26

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Chapter 27

1. *Device for Locating Faults in Wire Lines. Capacitance Bridge and Portable Locator*, Howard P. Corwith, OSRD 634, OEMsr-316, Project C-37, Western Union Telegraph Company, June 15, 1942. Div. 13-207.31-M2
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Chapter 28

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Chapter 29

1. *An Electrical Cancellation and Indicating System*, H. H. Beverage, OSRD 1408, OEMsr-748, Project C-67, RCA, Apr. 6, 1943. Div. 13-207.5-M1

Chapter 30

1. *Laying Communication Wires from Planes*, J. J. Gilbert, OSRD 1651, OEMsr-879, Project C-72, BTL, June 30, 1943. Div. 13-207.32-M2
2. *Trial of Wire Laying by Aircraft*, J. J. Gilbert, OSRD 1513, OEMsr-879, Project C-72, BTL, May 10, 1943. Div. 13-207.32-M1
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Chapter 31

1. *General Research in Field of Communication*, M. L. Almquist, OSRD 5741, OEMsr-1413, Project 13-104, BTL, Aug. 31, 1945. Div. 13-200-M1

Chapter 32

1. "Investigation of Engine Noise Problems," V. T. Callahan and D. F. Seacord, *Systems Engineering for Army Air Forces Communication*, Part II, Memorandum No. 6B, OSRD 1925, OEMsr-1018, Project C-79, BTL, Oct. 1, 1943. Div. 13-200.1-M2
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Chapter 33

1. *Investigation of Principles Underlying the Maximizing of Communication Intelligibility*, W. J. Cunningham, OSRD 6280, OEMsr-144, Project 13-110, Problem No. 4, Cruft Laboratory, Harvard University, Dec. 1, 1945. Div. 13-100-M2
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Chapter 34

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Chapter 35

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OSRD APPOINTEES

Division 13

Chief

C. B. Jolliffe (December 1942 to May 1945)
Haraden Pratt (May 1945 to May 1946)

Deputy Chief

K. C. Black

Technical Aides

J. L. Allison
C. F. Dalziel
J. F. McClean
A. F. Murray

Members

K. C. Black
O. E. Buckley
J. H. Dellinger
W. L. Everitt
G. C. Fick
R. H. George
C. H. G. Gray
A. Hazeltine
J. A. Hutchinson
C. M. Jansky
L. F. Jones
D. G. Little
R. K. Potter
H. Pratt
C. A. Priest
F. M. Ryan

Section Heads

Section 13.1	Direction Finding	L. F. Jones
Section 13.2	Radio Propagation Problems	J. H. Dellinger
Section 13.3	Speech Secrecy	R. K. Potter
Section 13.4	Special Communications Problems	C. A. Priest
Section 13.5	Precipitation Static	H. Pratt
Section 13.6	Miscellaneous Projects	D. G. Little

Consultants

L. V. Berkner
H. H. Beverage
E. D. Blodgett
D. G. Little
R. K. Potter

Interference Reduction Committee

K. C. Black
H. D. Doolittle
R. G. Fluharty
A. Hazeltine
J. C. R. Licklider
C. T. Morgan
A. F. Murray
S. S. Stevens
O. W. Towner

CONTRACT NUMBERS, CONTRACTORS, AND SUBJECTS OF CONTRACTS

Contract Number	*Name and Address of Contractor*	*Subject*	*Refer to Chapter*
NDCrc-75	Radio Corporation of America Camden, New Jersey	2-3,000 mc receivers, 2 IF amplifiers, 1 signal generator.	2
NDCrc-141	General Radio Company Boston, Massachusetts	UHF field intensity measuring equipment.	7
NDCrc-177	Western Electric Company, Incorporated New York, New York	RF generator.	2
NDCrc-88	Radio Corporation of America Camden, New Jersey	Aircraft facsimile.	7
Project C-9	National Bureau of Standards Washington, D. C.	Radio wave handbook.	6
NDCrc-191	Westinghouse Electric and Manufacturing Company Baltimore, Maryland	2-3,000 mc transmitters.	2
OEMsr-200	Carnegie Institution of Washington Washington, D. C.	Coordination of DF-ionospheric measurements.	6
OEMsr-227	Stanford University Stanford University, California	DF-ionospheric measurements.	6
OEMsr-92	Oregon State College Corvallis, Oregon	Precipitation static reduction.	3
OEMsr-378	Louisiana State University University, Louisiana	DF-ionospheric measurements.	6
OEMsr-32	Radio Corporation of America Camden, New Jersey	3,000 mc communication field research.	2
OEMsr-89	Farnsworth Television Company Fort Wayne, Indiana	Radio interference generator.	5
OEMsr-285	International Telephone and Radio Laboratories New York, New York	Radio interference generator.	5
OEMsr-49	Western Electric Company, Incorporated New York, New York	Panoramic receiver with moving screen indicator.	4
OEMsr-50	Western Electric Company, Incorporated New York, New York	Extra-high speed "flash" telegraphy.	7
OEMsr-120	Brush Development Company Cleveland, Ohio	Substitute for quartz crystals.	8
OEMsr-174	Jansky and Bailey Washington, D. C.	Mobile transmitter radio field survey.	1
OEMsr-225	Rensselaer Polytechnic Institute Troy, New York	Shielding for diathermy.	8
OEMsr-357	Western Electric Company, Incorporated New York, New York	Improved panoramic receiver.	4
OEMsr-316	Western Union Telegraph Company New York, New York	Device for locating faults in wire lines.	8
OEMsr-311	Western Electric Company, Incorporated New York, New York	Panoramic receiver for pulse signals.	4
OEMsr-420	Willard Storage Battery Company Cleveland, Ohio	Storage batteries for cold climates.	8
OEMsr-678	University of New Mexico Albuquerque, New Mexico	Precipitation static reduction.	3
OEMsr-442	Radio Corporation of America Camden, New Jersey	Voice communication on CM waves.	2
Project C-44	National Bureau of Standards Washington, D. C.	Supplement to radio wave handbook.	6
OEMsr-632	University of Puerto Rico Rio Piedras, Puerto Rico	Ionospheric measurement research.	6
OEMsr-573	Louisiana State University University, Louisiana	Ionospheric measurement research.	6
OEMsr-590	Stanford University Stanford University, California	Ionospheric measurement research.	6
OEMsr-558	Carnegie Institution of Washington Washington, D. C.	Ionospheric measurement research.	6
Project C2-49	National Bureau of Standards Washington, D. C.	Radio transmission measurements.	6
OEMsr-594	Carnegie Institution of Washington Washington, D. C.	Correlation of solar and geomagnetic observations with conditions of the ionosphere.	6

CONTRACT NUMBERS, CONTRACTORS, AND SUBJECTS OF CONTRACTS (*Continued*)

Contract Number	*Name and Address of Contractor*	*Subject*	*Refer to Chapter*
OEMsr-626	Western Electric Company, Incorporated New York, New York	Study of interference generation.	5
OEMsr-690	Radio Corporation of America Camden, New Jersey	Frequency stabilized master oscillator.	7
OEMsr-706	Radio Corporation of America Camden, New Jersey	Reconnaissance television.	7
OEMsr-679	Purdue Research Foundation Lafayette, Indiana	Study of effect of aircraft surface treatment on electrical charges.	3
OEMsr-748	Radio Corporation of America Camden, New Jersey	Electrical cancellation and indicating system.	8
OEMsr-893	United Air Lines Chicago, Illinois	Precipitation static research.	3
OEMsr-833	Brush Development Company Cleveland, Ohio	Sound recording on magnetic materials.	7
OEMsr-848	State College of Washington Pullman, Washington	Precipitation static exploration.	3
OEMsr-879	Western Electric Company, Incorporated New York, New York	Wire laying by aircraft.	8
OEMsr-1018	Western Electric Company, Incorporated New York, New York	Systems engineering for AAF communications.	1
OEMsr-1010	Jansky and Bailey Washington, D. C.	Measure of effect of hills and trees on radio propagation.	1
OEMsr-1086	Radio Corporation of America Camden, New Jersey	A survey of magnetic recording.	7
OEMsr-1413	Western Electric Company, Incorporated New York, New York	General research in the field of communications.	8
OEMsr-1441 Problem No. 2	Central Communications Research Cruft Laboratory, Harvard University Cambridge, Massachusetts	Study of FM versus AM.	1
OEMsr-1441 Problem No. 4	Central Communications Research Cruft Laboratory, Harvard University Cambridge, Massachusetts	Investigation of principles underlying maximizing communications intelligibility.	8
OEMsr-1441 Problem No. 6	Central Communications Research Cruft Laboratory, Harvard University Cambridge, Massachusetts	Aircraft transmitter antenna tuning network survey.	8
OEMsr-1441 Problem No. 8	Central Communications Research Cruft Laboratory, Harvard University Cambridge, Massachusetts	Measurement of magnetic properties of ferrite core materials.	8
OEMsr-1441 Problem No. 9	Central Communications Research Cruft Laboratory, Harvard University Cambridge, Massachusetts	Microwave survey	–
OEMsr-1475	General Electric Company Schenectady, New York	Investigation of radio noise generation in aircraft electrical machinery.	8
OEMsr-1478	University of Pennsylvania Philadelphia, Pennsylvania	Investigation of radio interference conditions on shipboard up to 1150 mc.	8

The projects listed below were transmitted to the Executive Secretary, NDRC, from the War or Navy Department through either the War Department Liaison Officer for NDRC or the Office of Research and Inventions (formerly the Coordinator of Research and Development), Navy Department.

Service Project Number	*Subject*	*Refer to Chapter*
AC-54	Systems engineering for AAF communications.	1
AC-54	Measure of effect of hills and trees on radio propagation.	1
AC-230.03	Study of FM versus AM.	1
AC-238.03	Investigation of radio noise generation in aircraft electrical machinery.	8
AC-238.07	Aircraft transmitter antenna tuning network survey.	8
SC-2	Precipitation static reduction.	3
SC-2	Precipitation static reduction.	3
SC-2	Study of effect of aircraft surface treatment on electrical charges.	3
SC-2	Precipitation static research.	3
SC-2	Precipitation static exploration.	3
SC-13	2-3,000 mc receivers, 2 IF amplifiers, 1 signal generator.	2
SC-13	3,000 mc communication field research.	2
SC-13	Voice communication on CM waves.	2
SC-15	Aircraft facsimile.	7
SC-16	UHF field intensity measuring equipment.	7
SC-18	Frequency stabilized master oscillator.	7
SC-19	Radio interference generator.	5
SC-19	Radio interference generator.	5
SC-19	Study of interference generation.	5
SC-20	Storage batteries for cold climates.	8
SC-23	Device for locating faults in wire lines.	8
SC-24	Substitute for quartz crystals.	8
SC-42	Wire laying by aircraft.	8
NA-110	Reconnaissance television.	7
NS-365	Investigation of principles underlying maximizing of communications intelligibility.	8
NS-368	Investigation of radio interference conditions on shipboard up to 1150 mc.	8

INDEX

The subject indexes of all STR volumes are combined in a master index printed in a separate volume. For access to the index volume consult the Army or Navy Agency listed on the reverse of the half-title page.

AAF communications systems, 3–6
  see also Communications systems
AFC
  see Automatic frequency control
Air Warning Service
  choice of frequency band, 4
  communications systems studies, 3
Aircraft antenna impedance studies, 200–201
Aircraft electric machinery noise, 193
Aircraft facsimile system, 141–143
Aircraft lightning strike hazard, 45
Aircraft static charge
  see Precipitation static studies
Aircraft surface treatment for static reduction, 51–52
Aluminum phosphate crystals for frequency control, 174
Amplitude modulation systems
  band space requirements, 13, 14
  comparison with f-m systems, 9
  crosstalk in multiplexed transmission, 13
  effect of mistuning, 8
  in directional microwave telephone transmitter, 36
  response to random noise, 8
AN/ARQ-9 radio set, scanning receiver research, 108–116
Antennas
  directive antennas for interference protection, 15
  effect of height on microwave propagation, 20
  effect on frequency stability of microwave transmitters, 22
  for directional microwave telephone, 34
  for mobile radiotelephone communications, 7
  for noise meters, 194
  for omnidirectional microwave telephone receiver, 27
  for radar pulse scanning receiver, 110
  impedance studies, 200–201
  ionosphere recording antennas, 132
  lightweight antenna mast, 4
  loading range for specific transmitters, 200–201
  site studies, 4
AN/TRC-5 multiplying method, 12
Atmospheric disturbances
  in mobile communications systems, 7
  ionosphere studies, 131–133
  precipitation static studies, 45–60
Attenuation, microwave
  in plywood, 26–27
  in rectangular guide at 9.8 cm, 25
  mutual inductance type attenuators, 139
  of radio waves by absorption, 132
  panoramic receiver input attenuator, 89
Audio limiters
  see Limiters
Aurora displays, effect on radio transmission, 133
Automatic frequency control
  cavity stabilization, 16
  effect on frequency stability, 158
  microwave transmitters, 23
Automatic volume control
  panoramic receivers, 66–67
  radar pulse scanning receiver, 112
Azimuth indicating system, radar pulse scanning receiver, 110

Balancers for noise reduction
  in a-m systems, 8
  in f-m systems, 9
  reduction of diathermy interference, 175
Band space requirements
  a-m systems, 12
  f-m systems, 8, 9, 11
  jamming systems, 119
  multiplex microwave transmission, 14
  pulsed emission, 12
  simplex microwave transmission, 13–14
Band-pass amplifiers, response analysis, 196–197
Batteries, low temperature, 181–182
BC-603 receivers, 9
Beam deflection converter receiver, 25
Beam deflection methods for panoramic indicators, 73, 78
Bendix Aviation Corporation, 51
Bendix charge dissipator, 48, 58
Blanking for panoramic reception, 82, 100
Block-and-squirter system for control of precipitation static, 59–60
Bolometer bridge, 31
Boonton Type 100-A *Q* meter, 198, 201
Bridge stabilized oscillator, 158
Bridges for wire fault location, 176–177
Butterfly circuit, 137–139
Buzzer noise for jamming speech signals, 119

C-25 jammer, 127
C-26 jammer, 124–127
C-59 frequency-stabilized master oscillator, 154–161
  circuit studies, 155–158
  humidity effects, 159
  importance of high Q, 159
  stability tests, 154
  temperature effects, 159
  tube studies, 154, 156
  tuning methods, 159–161
Camouflage paints, effect on aircraft static, 51–52
Cancellation and indicating system for complex waves, 183–190
  alarm and indicating system, 187–188
  approximation by rectangular pulses, 183–184
  complex wave generator, 184–185
  electrical cancellation, 188
  monitor oscilloscope, 187–188
  multiple station monitoring, 190
  panoramic reception, 188–189
  use as coding signal generator, 189–190
Carnegie Institution of Washington, field intensity and ionosphere studies, 131
Carrier frequency
  effect of doppler shift on bandwidth requirements, 13
  number of channels in a given band space, 13
  selection, 13
  stabilization, 13, 15, 16
Cathode-ray tubes as panoramic indicators
  amplitude vs. frequency diagrams, 78
  azimuth vs. frequency diagrams, 79
  choice of cathode ray screen type, 76
  facsimile type diagrams, 77
  for long scale 3- to 10-mc panoramic receiver, 99–100
  long persistence screens, 76
  moving screen tube, 77, 92–93
  recording radio telegraph signals, 77–78, 83–85
  resolution, 76–77
  signal presence vs. frequency diagrams, 73–76
Cavity for 1290-CT r-f oscillator tube, 37
Cavity monitor, r-f generator, 38–39
Cavity oscillator for directional telephone, 35
Cavity stabilization scheme for afc, 16
CFVD modulation (continuous frequency, variable duration) pulse system, 22, 23, 36
  r-f generator, 41
Charge dissipators, aircraft static
  flame discharge, 48
  radioactive cups, 48
  trailed wire type, 47, 48, 55–56
Charge indicator, aircraft static, 56
Charging rates of aircraft camouflage paints, 51–52
Clippers
  see Limiters
Coding signal generator, electrical cancellation system, 189–190
Colpitts oscillator circuit, frequency stability studies, 155–156
Communication uses for panoramic receivers, 86
Communications systems
  see also Microwave communications systems, design factors

  directional microwave telephone, 32–36
  facsimile system, 141–143
  flash telegraphy systems, 144–153
  mobile radio telephone communications, 6–7
  omnidirectional microwave telephone, 27–32
Complex impedance
  see Impedance
Complex waveform approximation using rectangular pulses, 183
Complex-wave generator, commutator type, 184
Compression amplifier, moving-screen panoramic receiver, 91, 95
Continuous waves for communications, advantages, 15
Core materials, magnetic properties
  see Magnetic properties of ferrite core materials
Corona discharge from planes
  current measurements, 58–59
  voltage induction effects, 47
CR-301 microwave receiver, 25
Cross-band signaling, 14–16
Crosstalk in multiplexed transmission, 13
Crystal control for ultra high frequencies, 13
Crystal converter microwave receivers
  disadvantages, 21–22
  frequency stability, 24
  input noise characteristics, 24
  sensitivity, 24
Crystals, for frequency control, synthetics, 173–174
CuZn ferrite, magnetic permeability, 199
C-W Morse telegraph operation at v-h-f, 4

D 160127 coaxial line tube, 24, 139
Dampers, for reducing precipitation static, 47
Deaf persons aided by visible speech system, 177
Diathermy equipment shielding, 175
Dielectric constant of plywood, 26
Direction finding, panoramic receivers, 87
Directional microwave telephone
  antenna system, 34
  field tests, 36
  power supply, 35
  requirements, 32–34
  r-f unit, 35
  transmitter, 36
Discharge wires for static dissipation, 48
Discriminator, f-m systems, effect on pulse interference, 9, 10
Dry cells for low temperature use, 181–182
Duty cycle, panoramic receiver, 109
Dynamotor noise studies, 193

Electrolytic recorder, facsimile, 141
Electron gun, television, 162, 165
Electrostatic charging rates of aircraft in flight, 58–59
ENR (excess noise ratio)
  definition, 31–32
  directional microwave telephone receiver, 35
  omnidirectional microwave telephone receiver, 31
ENSI (equivalent noise sideband input)
  definition, 21
  of crystal receivers, 21
  microwave receivers, 24–25
Erasing magnetic recordings, 144
E-region ionization, effect on radio transmission, 133

Facsimile system, aircraft, 141–143
Facsimile-type diagrams, panoramic receivers, 77
Fault location in wire lines, 176
Ferrite, magnetic cores, 198–199
  permeability, 198
  specific loss factor, 198
Fiberglas, use in floating insulated wire, 192
Fibrite as storage battery insulator, 181
Field intensity studies
  ionosphere studies, 131–132
  u-h-f measuring equipment, 137–140
Field wire, laying by airplane over jungle, 191
Flame discharger for precipitation static reduction, 48
Flash telegraphy system, 144–153
  advantages, 144
  booster amplifier, 148
  distortion from tape recording, 150–152
  flasher design, 147–148
  magnetic tape recording system, 144–145, 152–153
  radio circuit requirements, 153
  starting pulse problems, 146-147
  tape drive mechanism, 148
  teletypewriter and perforated tape methods, 145
Floating wire cable on water, methods, 192
F-M systems
  see Frequency modulation systems
Fortisan, use for floating insulated wire, 192
Free space field, field intensity studies, 7
Frequency allocation
  Air Warning Service radio circuits, 4
  mobile radiotelephone communication, 6
  panoramic receivers, 64–65
Frequency control, automatic, 16, 23, 158
Frequency determination, panoramic receivers, 81–82
  beating oscillators, 82
  calibrated signal generator, 81
  electrical markings, 81
  fixed frequency markings, 81
Frequency difference compensator, microwave telephone, 36
Frequency modulation systems, 8-10, 23
  band space requirements, 12–14
  comparison with a-m, 9–10
  crosstalk in multiplexed transmission, 13
  jamming, 125, 126
  narrow band and wide band systems, 8, 9
  overmodulation, 9
  response to random noise, 8–10
  signal-to-noise ratio improvement, 24
  sweep oscillators for panoramic reception, 64, 106
Frequencies for communication systems
  see Band space requirements
Frequency spectrum of precipitation static, 48
Frequency stability
  circuit studies, 155–157
  humidity effects, 159
  importance of high Q, 159
  microwave telephone oscillator, 32
  microwave transmitter, 22–25
  stability tests, 154
  temperature effects, 159
  tube studies, 154–156
  2-20 mc oscillators, 154–159
Frequency-division multiplexing, 12–13
Frost crystals, effect on airplane static charge, 51–52

Geomagnetic phenomena, ionosphere studies, 131–133
Glass wool as storage battery insulator, 181
Glide-path indication, use of panoramic receivers, 87
G-meter for vertical acceleration measurements, 50
Gradient indicator, aircraft, electrostatic, 56
Grounding, effect on shielded room, 175
Gum-bichromate process for television tubes, 164

Hartley oscillator
  for frequency stabilized 2-20 mc oscillator, 154
  frequency stability studies, 156
  in moving-screen panoramic receiver, 90
Heterodyne methods in panoramic reception, 65
Hills, obstructions to radio propagation, 5, 6
Horn antennas for microwave use, 25
Humidity effect on oscillator stability, 159

Ignition disturbances in mobile communications system, 7
Impedance
  aircraft antennas, 200–201
  effect of weather on antenna matching, 22
  measuring equipment, 201

  - plywood filled waveguide, 26
- Impulse noise for jamming speech signals, 119
- Indicators for panoramic receivers
  - amplitude vs. frequency diagram, 78
  - azimuth vs. frequency diagrams, 79
  - choice of screen type, 76
  - facsimile type diagram, 77
  - long scale indicating equipment, 100–102
  - mechanical indicators, 79
  - moving screen tube, 77
  - recording radio telegraph signals, 77-78
  - repetition rate, 76
  - resolution, 76–77
  - signal alarms, 80
  - signal presence vs. frequency diagrams, 73–76
  - 350-750 mc receiver, 109–110
- Information center communications systems, 3
- Infrared rays for wiping out CRO traces, 95
- Instrument landing, use of panoramic receivers, 87
- Insulation, low temperature storage batteries, 181
- Intelligibility tests, speech transmission jamming, 119–120
- Interference
  - *see also* Jamming studies
  - effect on omnidirectional microwave telephone, 32
  - pulse interference, 8, 9
  - random noise in a-m systems, 8
  - random noise in f-m systems, 8–10
  - reduction in AAF communications systems, 4, 195
  - shielding for diathermy equipment, 175
- Interference Reduction Committee, 195
- Interservice Radio Propagation Laboratory (IRPL), 131
- Inverse vacuum-tube voltmeter, 49
- Ionosphere studies, 131–133
  - apparatus, 132–133
  - field strength studies, 131–132
  - solar and geomagnetic observation, 133

- Jamming, 117
  - C-25 jammer, 127
  - C-26 jammer, 124–127
  - criteria for successful jamming, 123–124
  - effect on a-m and f-m receivers, 9
  - efficiency of noise limiters against high intensity pulses, 126
  - field test results, 126–127
  - form and degree of modulation, 124, 126, 127
  - fundamental masking studies, 122–123
  - interference with f-m transmission, 125
  - interference with speech transmission, 119–120
  - interference with telegraph transmission, 120-121, 123
  - military considerations, 124
  - protection against jamming, 15
  - resistance noise generation, 121
  - tuning to enemy frequency, 124–125, 127
  - use of panoramic receivers, 87
- Jungle terrain, laying field wire by airplane, 191

- Klystron transmitter, 14, 19, 23–24

- Latex sponge as storage battery insulator, 181
- Laying field wire by airplane, 191
- Leland Stanford University, field intensity and ionosphere studies, 131
- Lightning strikes, aircraft, 45
  - gradient warning indicator, 56
- Lighthouse tubes
  - in directional microwave telephone oscillator, 35
  - limitations for butterfly tuning circuits, 139
- Limiters
  - effect on jamming efficiency, 126
  - for microwave telephone transmitter, 30
  - for noise reduction, 8–10, 15
  - for panoramic receivers, 66–67
  - for precipitation static reduction, 47
  - in a-m systems, 8, 9–10
  - in band-pass amplifiers, 196–197
  - in f-m systems, 9
- Line mosaic pick-up tube, 162–165
  - electron gun, 162, 165
  - mosaic types, 163–164
  - optical system, 162
  - test results, 165
  - tube construction, 164–165
- Lineman's portable open circuit locator, 177–178
- Line-of-sight path for microwave propagation, 19, 21
- Loading range measurements, r-f transmitter, 200, 201
- Long-scale panoramic receiver, 98–104
  - balanced modulators, 103
  - beam modulation method of indication, 99
  - blanking out recognized stations, 100
  - cathode-ray tube, 99–100
  - length and shape of scale, 98–99
  - method of obtaining spiral trace, 99
  - modulating amplifier, 102
  - operation of indicating equipment, 100–102
  - phase shift circuits, 103
  - scanning filters, 99–100
  - scanning rate, 100
  - synchronization amplifier, 104
- Loss factor, magnetic core materials, 198, 199
- Louisiana State University, field intensity and ionosphere studies, 131
- "Magic T" waveguide section, for cross-band operation, 16
- Magnetic disturbances
  - effect on radio transmission, 132
  - radio propagation ionosphere studies, 131–133
- Magnetic properties of ferrite cores, 198–199
  - permeability, 198, 199
  - Q, 198–199
  - specific loss factor, 198
- Magnetic recording
  - a-c erase, 144
  - a-c recording, 145
  - advantages, 166
  - blanking out recognized stations in panoramic reception, 82
  - circuits, 152
  - d-c erase, 144
  - d-c recording, 144–145
  - distortion, 150–152
  - drive mechanism for tape, 148
  - flash telegraphy system, 144–145, 152–153
  - pocket-sized wire recorder, 166–169
  - recording without bias, 146–147
  - reproducing system, 152–153
- Magnetron
  - in omnidirectional microwave telephone transmitter, 27
  - output power, 14
- Marking arrangements, panoramic receivers
  - broad band 0.1- to 30-mc scanning receiver, 107
  - moving screen receiver, 94, 95
- Masking studies, 122-123
  - by clicks and pulses, 122
  - by noise, 122
  - code signals, 123
  - pure tones, 122
  - speech intelligibility, in the presence of noise, 120
- Mechanical indicators for panoramic receivers, 79
- Micromax line-drawing type of recorder, 132
- Micrometer-tuned wavemeter, 38
- Microphone, single-button carbon, 28
- Microwave communications systems, design factors, 11–43
  - antennas, 25, 34
  - cross-band signaling, 14, 16
  - i-f power, 14
  - multiplexing of signals, 12–13
  - omnidirectional microwave telephone, 27–32
  - propagation studies over land and sea water, 19
  - protection from interference and jamming, 15–16
  - receivers, 24, 31–32
  - recommendations for universal set, 15–16
  - r-f generator, 37–43
  - selection and stabilization of carrier frequency, 13, 16
  - spectrum widths, 11–12

[redacted]

  transmitters, 22–24, 28–31, 36
  types of operation, 11
  use of band space, 13–14
  waveguides, 25
Microwave frequency discriminator for stabilization, 16
Mistuning, effect on receiver performance, 8, 9
MnZn ferrite, magnetic permeability, 199
Mobile microwave communications system
  see Directional microwave telephone
Mobile radiotelephone communication, 6–7
  antenna characteristics, 7
  choice of frequency, 6, 7
  field intensity studies, 7
Modulating amplifier
  long-scale 3- to 10-mc panoramic receiver, 102
  moving screen panoramic receiver, 92
Modulation
  see also Amplitude modulation systems; Frequency modulation systems
  cathode modulation of directional microwave telephone transmitter, 36
  CFVD system, 41
  overmodulation effect on f-m systems, 9
  types of pulse modulation, 11
Monitor cavity, r-f generator, 41–42
Monitor circuit use of electrical cancellation system, 190
Monitor oscilloscope for electrical cancellation system, 187
Morse operation, at v-h-f, c-w, 4
Mosaic types, line mosaic television pick-up tube, 163–165
Motor noise studies, radio, 193
Mountainous terrain, propagation studies, 5–6, 7
Moving-screen panoramic receiver, 77, 88-97
  advantages, 88
  amplifiers, 92–95
  band-pass filters, 90, 91
  cathode-ray tube, 92–93
  design principles, 88–89
  input attenuator, 89
  marking arrangement, 94, 95
  modulating amplifier, 92–93
  power supplies, 94
  pulse reception studies, 96–97
  reproduction and reading of telegraph code, 88
  r-f and modulator circuits, 90
  scanning oscillator, 90, 94–95
  three dimensional pattern, 88
Multipath transmission, f-m systems, 9
Multiple responses in panoramic reception, 65
Multiple selection circuits, 86
Multiplex operation, microwave communications system, 11–14
  band space requirements, 14
  crosstalk problem, 13
  c-w carrier frequency stabilization, 16
  frequency division multiplexing, 12 13
  time division multiplexing, 12–13
Multivibrator circuit for pulse generator, 42

Narrow-band scanning, panoramic reception, 82–83
National Bureau of Standards, field intensity and ionosphere studies, 131
Navigation, using panoramic radio receivers, 87
Network analysis, band-pass, 196
Network analysis, use of complex-wave generator, 190
Noise for jamming speech transmission, 119–120
Noise spectrum from precipitation static, 47, 48
Noise limiters
  see Limiters
Noise studies
  see also Jamming
  aircraft electrical machinery, 193
  meter studies, 193–194
  precipitation static, 45–60
  random noise, 8–10
Nomograph for optimum filter bandwidth, 70
Nylon for floating insulated wire, 192

Omnidirectional microwave telephone, 27–32
  antennas, 27, 32
  effect of interference, 32
  probable range with one set in airplane, 32
  receiver, 31–32
  transmitter, 27–31
116 mc communication systems
  effect of increasing antenna height, 6
  for transmission over broken country, 5
  mobile radiotelephone communication, 7
Open circuits in lines, location
  capacitance bridge method, 176–177
  errors due to resistance and leakage, 177
  portable locator, 177–180
Oscillators
  300 to 1,000 mc, 137–139
  410 to 690 mc, 111
  1,000 to 3,000 mc, 139
  2,000 to 3,000-mc r-f generator, 37–43
  C-59 frequency stabilized oscillator, 154, 159–161
  frequency stability studies, 154–161
  in directional microwave telephone, 35
  sweep oscillators in panoramic receivers, 64–65, 111
Oscilloscopes
  see also Cathode-ray tubes as panoramic indicators
  monitor for electrical cancellation system, 187
Overmodulation effect in f-m systems, 9
Owen bridge for measuring inductance, 198

Paillard Bolex 16-mm camera, 50
Painted surfaces, precipitation static accumulation, 56
PAM (pulse amplitude modulated) emission, 11
Panoramic facsimile recorder, 83–85
Panoramic receivers, 61–116
  automatic control of signal intensity, 66–67
  blanking out recognized stations, 82
  broad band 0.1- to 30-mc scanning receiver, 104–108
  comparison with ordinary radio reception, 63–64
  electrical cancellation and indicating system, 188–189
  frequency allocations, 64–65
  frequency determination, 80–82
  long-scale model for 3 to 10 mc range, 98–107
  moving-screen indicator, 88–97
  narrow band scanning, 82
  panoramic indicators, 72–80
  radar pulse reception, 85–86
  range expansion, 83
  receivers without frequency sweep, 86
  recording, 83–85
  research recommendations, 107, 114, 115
  resolution, 76–77
  sensitivity, 80
  scanning filter design and performance, 67–72
  350-750 mc scanning receiver, 108–114
  tuning jammer to enemy frequency, 124–125
  uses, 86–87
Parallel traces for panoramic indicator, 74–76
Perforated tape methods of flash telegraphy, 145
Permeability, magnetic
  formula, 198
  of ferrite materials, 199
PFM (pulse frequency modulation) emission, definition, 11
Pickup tube for reconnaissance television
  see Line mosaic pickup tube
π-type attentuator for panoramic receiver input, 89
PLM (pulse length modulated) emission, definition, 11
Plywood antennas
  dielectric constant, 27
  microwave losses, 26–27
PNM (pulse number modulated) emission, definition, 11
Polarization
  horizontal vs. vertical for broken country, 5

  transmission over salt water, 19–20
Polyethelene for floating insulated wire, 192
Portable wire recorder, 166–169
Power dividing wave guide sections, 16
Power measurement
  adjustable load method measurements for r-f transmitter, 200
  microwave standing wave detector, 27
Power supply
  directional microwave telephone, 35
  in moving screen panoramic indicator, 94
  requirements for microwave communications systems, 14
  r-f generator, 43
Power tubes for microwave systems
  klystron, 14
  magnetron, 14
PPM (pulse position modulated) emission
  advantages, 15–16
  definition, 11
Precipitation static studies, 45–60
  airplanes for test flights, 54
  apparatus, 47, 49–50, 54–55, 58
  block-and-squirter discharge system, 59–60
  causes of static, 57
  effect of aircraft surface treatment, 51–52
  flight tests, 56, 58–59
  flame discharger, 48
  noise frequency spectrum, 47, 48
  nonlinear elements for receiver circuits, 47
  trailed wire charge dissipators, 47–48, 54, 55–56, 59
  voltmeter for high voltage measurements, 49
Propagation studies, radio waves
  choice of transmitter and receiver locations, 6
  effect of antenna heights, 6
  effect of polarization, 19–20
  effect of type of terrain, 5–6, 7
  ionosphere studies, 131–133
  over land at microwave frequencies, 20–21
  radiotelephone communication between mobile units, 6–7
Psycho-Acoustic Laboratory, Harvard University, 8
PTM (pulse time modulated) emission, definition, 11
Pulse amplitude modulated (PAM) emission, definition, 11
Pulse control, microwave telephone transmitter, 30
Pulse frequency modulated (PFM) emission, definition, 11
Pulse generator, r-f generator, 42–43
Pulse jamming
  in a-m systems, 8, 9
  in f-m systems, 9
Pulse length modulated (PLM) emission, definition, 11
Pulse number modulated (PNM) emission, definition, 11
Pulse position modulated (PPM) emission, definition, 11
Pulse reception
  by panoramic receivers, 85–86
  moving screen panoramic receiver, 96–97
  pulse scanning receiver, 108–116
Pulse scanning receiver, 350-750 mc, 108–116
  alarm and motor control, 112–115
  antenna, 110
  a-v-c circuit, 113
  converter, 110
  detector, 113
  duty cycle, 109
  energy distribution in radar pulses, 108
  filter width, 108
  i-f amplifiers, 113
  indicating systems, 109, 110
  monitor and sweep circuits, 115
  oscillator, 111–113
  receiver characteristics, 115
  recommendations for future research, 115–116
  signal-to-noise ratio, 109
  video amplifier, 113
Pulse time modulated (PTM) emission, definition, 11
Pulse transmitter, omnidirectional microwave telephone, 28–31
  audio limiter, 30
  circuit operation, 30–31
  magnetron, 28–29
  output circuit, 31
  pulse control, 30
  wavemeter, 31
Pulsed wave emission, 11–15
  band space requirements, 11–12, 13–14
  bandwidth formula, 12
  for relay and trunkline service, 15
  time division multiplexing, 12–13
  types of modulation, 11
Pulse-modulated telephone generator
  see R-f generator, 2,000 to 3,000 mc
Purdue Research Foundation, 51

Q
  of frequency stabilized oscillator, 159–160
  of magnetic core materials, 198
Quartz crystal substitutes for frequency control, 173–174

Radar pulse reception, panoramic receivers, 85–86
Radar pulse scanning receiver
  see Pulse scanning receiver, 350-750 mc
Radio navigation, use of panoramic receivers, 87
Radio propagation
  see Propagation studies, radio waves
Radio transmission, ionosphere studies, 131–135
Radioactive cups for precipitation static dissipation, 48
Radiotelephone communication
  antenna characteristics, 7
  between mobile units, 6–7
  choice of frequency, 6, 7
  directional microwave telephone, 32–36
  field intensity studies, 7
  omnidirectional microwave telephone, 27–32
  studies of effective speech jamming, 119–120
Rain, electrostatic charge from, 58
Random noise, 8–10
Range, maximum
  for f-m systems, 9
  for microwave communications, 14
  microwave propagation studies over salt water, 19
  of directional microwave telephone, 34, 36
  of omnidirectional microwave telephone system, 32
Range expansion, panoramic reception, 82
RCD-23 microwave receiver, 25
Reactance
  see Impedance
Receivers, radio
  see also Panoramic receivers
  0.1 to 30 mc, 104–108
  3 to 10 mc, 99–100
  300 to 1,000 mc, 139
  350 to 750 mc, 108–116
  1,000 to 3,000 mc, 139
  antenna for microwave receivers, 32
  apparatus for ionosphere field and intensity studies, 132
  CR-301 beam deflection converter receiver, 25
  effect of mistuning, 8
  for omnidirectional telephone system, 31–32
  frequency stability, 24, 25, 32
  input noise characteristics, 24, 25
  location for radio transmission over broken country, 6
  microwave crystal receivers, 21, 24
  RCD-23 receiver, 24–25
Recommendations for future research
  camouflage paint binder for reduction of precipitation static, 52
  charging rate of semiconducting rubber, 52
  charging rates measurements on aircraft wing surface, 52
  impedance measuring equipment, 201
  noise measurement studies, 195
  panoramic reception, 107, 115–116
  quartz crystal substitutes for frequency control, 173–174
  two dimensional mosaic television pickup tube, 165
  universal microwave communication equipment, 15–16
Recorders
  for aircraft facsimile system, 141

  for ionosphere field intensity studies, 132
Recording, magnetic tape
  see Magnetic recording
Recording head, pocket sized wire recorder, 169
Recording radio telegraph code from panoramic indicator, 77–78, 83–85
Refraction studies, microwave, 21
Rejection circuits for panoramic reception, 82
Repetition rates
  for frequency diagrams, panoramic indicators, 76
  for panoramic facsimile recorder, 84
Resistance
  see Impedance
Resistance noise
  for jamming speech signals, 119, 120, 121
  for jamming telegraph, 120–121
  interference criteria, 123
  methods of generation, 121
Resolution in panoramic receivers, 76–77
R-F generator, 2,000 to 3,000 mc, 37–43
  automatic tuning, 38
  CFVD modulation, 41
  magnetic focusing, 37
  monitor cavity, 38–39
  oscillator tube, 37
  power supply, 43
  pulse generator, 42–43
Rochelle salt crystals for frequency control, 173
Runway localizing and marking, use of panoramic receivers, 87

Sawtooth generator for panoramic indicator, 93–94
Scanning filters, panoramic receivers, 67–72
  analogy with quasi-stationary filter response, 68–70
  bandwidth and resolution for equal-level signals, 69–70
  bandwidth formula, 108
  desirable characteristics, 72
  effect of level difference, 70–71
  effect on resolution, 68–69
  general design considerations, 67–68
  in long scale 3- to 10-mc receiver, 99–100
  location of scanning filter, 72
  radar pulse reception requirements, 85–86
  use of one filter at two or more scanning speeds, 72
Scanning oscillators in panoramic receivers
  see Sweep oscillators, panoramic receivers
Scanning receiver, 0.1- to 30-mc, 104–107
  see also Panoramic receivers
  beating oscillators, 106
  f-m oscillator, 106
  f-m tripler and amplifier receiver, 106
  input filters, 105
  modulators, 106
  monitoring receiver and marker, 107
  r-f amplifier, 106
SCR-508 f-m transmitter-receiver equipment, 9
Secrecy, radio
  advantages of flash telegraphy, 144
  use of 116 mc for mobile radiotelephone communications, 7
Selection circuits, panoramic reception for receivers without frequency sweep, 86
  frequency allocations, 65–66
Sensitivity, microwave receivers, 24
Sensitivity, panoramic receivers, 80–81
  radar pulse reception, 86
  radar pulse scanning receiver, 109
Shielding diathermy equipment, 175
Shock excitation, effect on radio receivers, 47
Signal amplitude selection circuits, 15
Signal generator, coding, 189–190
Signal generators
  see Oscillators
Signal intensity control, panoramic receivers, 66–67
Signal strength studies, microwave propagation, 21
Signal-to-noise ratio, 119
  see also Sensitivity, microwave receivers; Sensitivity, panoramic receivers
  advantages of f-m, 8, 15
  effect of i-f bandwidth in a-m systems, 8
  for radar pulse reception, 86, 109
  improvement using f-m in klystron transmitter, 24
  improvement with long persistence screen, 76
  scanning filters for panoramic receivers, 71
6C4 tube, frequency stability studies, 156
Slaytor Electronics Corporation, static discharger, 56
Slit-aperture electron gun, 162, 164
Snow crystals, effect on airplane static charge, 51, 58
Solar phenomena, ionosphere studies, 131, 132, 133
Solar-activity cycles
  effect on radio transmission, 133
Specific loss factor, magnetic core materials, 198, 199
Spectrograms of radio noise, 119
Spectrum widths
  see Band space requirements
Speech transmission, effective jamming
  see also Jamming
  continuous vs. interrupted jamming, 120
  interference criteria, 123
  noise spectrogram, 119
  required frequency spectrum, 119
  required noise-to-signal ratios, 119
  speech mixtures for jamming, 119
Spiral trace, panoramic receiver indication, 99
Spiral trace, signal presence diagrams, panoramic indicator, 74
"Spiral-four" audio-frequency multiplexing, 12
Square wave approximation of complex waveforms, 183–184
Stabilization of carrier frequency, 16
Stabilization of microwave carrier frequency, 13
Stagger-tuned circuits, 196–197
Standing wave measurements, 26, 27
Stanford University, field intensity and ionosphere studies, 131
Static
  see Precipitation static studies
Storage batteries for low temperature use, 181–182
Storm forecasting from ionosphere studies, 131–133
Sun-spot cycles, effect on radio transmission, 133
Supersonic-frequency multiplexing, 12
Sweep oscillators, panoramic receivers, 64–65
  electronically controlled oscillators, 64–65
  for moving-screen receiver, 90
  mechanically controlled oscillators, 64
  use of an RC oscillator, 94–95

Tape recording
  see Magnetic recording
Telegraph transmission
  flash telegraphy system, 144–153
  interference criteria, 123
  jamming studies, 120–121
  masking studies, 123
  reproduction and reading code from panoramic receiver indicator, 77–78, 83–85, 88
  single-channel teletypewriter system, 4
Telephone systems, radar
  directional microwave telephone, 32–36
  omnidirectional microwave telephone, 27–32
  speech jamming studies, 119–120
Telephone wire laying from airplanes, 4
Teletypewriter method of flash telegraphy, 145
Teletypewriter system, single channel, 4
Television, line mosaic pick-up tube, 162–165
Temperature effects on frequency stability of oscillator circuits, 32, 159
350 to 750 mc pulse scanning receiver
  see Pulse scanning receiver, 350-750 mc
"Three-gap" oscillator tube, 37
3,000 mc communication systems, design

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  - *see* Microwave communication systems, design factors
- Threshold of understandability of speech in the presence of continuous and interrupted noise, 120
- Thunderclouds indicator, 56, 59
- Time-division multiplexing, 12–13
- Trailed wire charge dissipators, 47–48, 54–56, 59
- Transmitters, microwave, 22–24
  - automatic frequency control, 22–23
  - CFVD modulator, 22, 23
  - effect of ambient temperature, 22
  - frequency stability, 22–23
  - in directional microwave telephone, 36
  - in omnidirectional telephone system, 28–31
  - loading and tuning range of aircraft antennas, 200–201
  - location for transmission over broken country, 5, 6
  - push-to-talk circuit, 24
  - r-f power output measurements, 200
- Trap circuits for blanking in panoramic reception, 82
- Trees obstructing wave propagation, 5, 22
- Triboelectric series
  - camouflage paints, 52
  - snow and frost crystals, 52
- Triple heterodyne circuit, 104, 106
- Tuning systems
  - adjustable load method, 200
  - butterfly circuit, u-h-f oscillators, 137–138, 139
  - C-26 jammer, 124–125
  - dummy antenna method of testing, 200, 201
  - frequency stabilized 2-20 mc oscillator, 159–161
  - r-f generator, 37–42

- U-H-F field intensity measuring equipment, 137–140
  - 300-1000 mc oscillator, 137–139
  - 1000-3000 mc oscillator, 139
  - receivers, 140
- University of Puerto Rico, field intensity and ionosphere studies, 131
- Vario-losser circuits, 67
- Velocity-modulated r-f generator, 2000-3000 mc, 37–43
- V-H-F frequency band
  - antenna site selection studies, 4
  - c-w Morse operation, 4
  - effect of obstructions on radio propagation, 4–6
  - for Air Warning Service radio circuits, 4
  - reduction of radio interference, 4
- Video amplifier, radar pulse scanning receiver, 112
- Visible speech system for aiding the deaf in learning to talk, 77
- Washington State College, 57
- Wave generator for complex waveforms, 184
- Wave propagation
  - *see* Propagation studies, radio waves
- Waveguide
  - curvature of bends for negligible reflection, 25
  - for microwave transmitter antenna feed, 25
  - loss in rectangular guide at 9.8 cm, 25
  - use of "magic T" sections in cross-band signaling, 16
- Wavemeter
  - in microwave telephone transmitter, 31
  - in r-f generator tuning system, 38
- Weather, effect on microwave transmission, 21, 22
- Wire, floating insulated, 192
- Wire, trailing, as charge dissipator for precipitation static, 47–48, 54–59
- Wire laying by airplane over jungle terrain, 4, 191
- Wire lines, fault location, 176–180
- Wire recorder, pocket sized, 166–169
  - d-c erase and bias, 167
  - drive mechanism, 166–167, 169
  - play back amplifier, 169
  - wire details, 167–169

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